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INTERCEPTION OF LPI RADAR SIGNALS (U)

by

Jim P.Y. Lee

DEFENCE RESEARCH ESTABLISHMENT OTTAWA
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Jim P.Y. Lee
Radar ESM Section
Electronic Warfare Division

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ABSTRACT

Most current radars are designed to transmit short duration pulses with relatively high peak power. These radars can be detected easily by the use of relatively modest EW intercept receivers. Three radar functions, namely search, anti–ship missile (ASM) seeker and navigation, are examined in this report to evaluate the effectiveness of potential low probability of intercept (LPI) techniques, such as waveform coding, antenna profile control and power management, that a radar may employ against current EW receivers. The general conclusion is that it is possible to design a LPI radar which is effective against current intercept EW receivers. LPI operation is most easily achieved at close ranges and against a target with a large radar cross section. The general system sensitivity requirement for the detection of current and projected LPI radars is found to be on the order of – 100 dBmi which cannot be met by current EW receivers. Finally, three potential LPI receiver architectures, using channelized, superhet and acousto–optic receivers with narrow RF and video bandwidths are discussed. They have shown some potential in terms of providing the sensitivity and capability in an environment where both conventional and LPI signals are present.

RESUME

La plupart des radars courants sont conçus pour transmettre des impulsions courtes de puissance maximale élevée. Ces radars se détectent facilement par des récepteurs de guerre électronique (GE) relativement simples. L'efficacité potentielle de techniques à faible probabilité d'interception (FPI) (codage d'onde, contrôle du profil du faisceau de l'antenne et gestion de la puissance des émissions) qui sont employées par les radars contre des récepteurs de GE courants, est évaluéee pour trois fonctions d'un radar soit la veille, la détection des missiles anti–navires et la navigation. La conclusion générale est qu'il est possible de concevoir un radar à FPI efficace contre les récepteurs de GE courants. Les opérations à FPI sont plus facilement réussies à courte distance et contre des cibles ayant une grande section efficace. La sensibilité nécessaire pour la détection des radars à FPI d'aujourdhui et de demain est de l'ordre de – 100 dBmi. Les récepteurs courants de GE sont incapables de telles performances. Finalement, trois architectures de récepteurs à FPI à bande passante RF et vidéo étroites, soient les récepteurs multibandes, superhétérodynes et acousto–optiques sont décrits. Leur sensibilité et leur capacité ont démontré un certain potentiel pour un environnement contenant des signaux conventionnels et à FPI.

EXECUTIVE SUMMARY

Most current radars are designed to transmit short duration pulses with relatively high peak power. These radars can be detected easily by the use of relatively modest EW intercept receivers. The intercept of radar transmissions ultimately leads to vulnerability through the use of either antiradiation missiles or ECM. However by using low probability of intercept (LPI) techniques, it is possible to design a LPI radar which is effective against current intercept EW receivers.

There are a number of LPI techniques a radar can employ. These may include low sidelobe antennas, infrequent scanning, power control when tracking a closing target (as range is reduced, the radar power is also reduced), making use of waveform coding to provide transmitting duty cycles approaching one (to reduce peak power while maintaining the required average power) and using frequency hopping to force the interceptor to consider more of the spectrum in attempting to characterize the radar.

In this report, an analysis is presented on current and projected LPI radar signals and the detection of these signals by EW receivers. The analysis starts with an introduction on the difference in detection between a radar receiver and an EW receiver. A radar receiver is designed to exploit the coherent integration gain of matched filters and the incoherent integration gain by integrating a number of pulses. On the other hand, current EW receivers are designed to cover a much broader RF bandwidth and to detect the shortest anticipated radar pulses and the resultant equivalent noise bandwidth (B_i) can be quite large. As a consequence, there is a mismatch between the radar transmitter waveform and an EW receiver. The relative mismatch is given by the time–bandwidth factor (τB_i) and τ is the duration of the radar pulse. This time–bandwidth factor is quite large for some current wide–open EW receivers. Despite this mismatch, the EW receiver has the range advantage due to one–way propagation loss. In addition, most current radars transmit short duration pulses with relatively high peak power. As a result, most current radars can be detected easily by the use of current EW receivers.

To make a radar LPI in which the radar cannot be intercepted beyond the range at which it can detect targets itself, a radar designer can maximize the mismatch further by increasing the duration of the signal. This can be carried out by employing signal waveforms in which the range resolution of the radar is recovered while the transmitted peak power can be reduced. As a result, LPI signals are expected to be of long duration and thus higher duty cycles. The EW receiver designer can also respond by minimizing B_i to match these LPI waveforms. However, it is difficult to build an EW receiver which can meet both the requirements of having a small equivalent noise bandwidth and be able to detect signals over a wide instantaneous RF bandwidth.

The ratio of the radar detection range to the EW intercept receiver detection range is derived. The detection range of three typical radars for the functions of (a) search, (b) anti–ship missile (ASM) seeker and (c) navigation, is then examined against the detection range of three typical EW receivers. The purpose is to evaluate the effectiveness of potential LPI techniques, such as waveform coding, antenna profile control and power management, that a LPI radar may employ against conventional EW receivers. It is shown that LPI operation is most easily achieved at close ranges only. In the search function, the range is usually quite large and the target size can be small. As a result, it is very difficult to design a radar LPI against conventional EW receivers when the mainbeam is intercepted. A combination of antenna sidelobe control and waveform coding are essential for LPI operation when the interceptor is located in the sidelobes. For the ASM seeker function the target size is relatively larger and the range is reduced when tracking a closing target. As a result, the techniques of power control and waveform coding can be effective for LPI operation. However, the complexity ,cost and space will probably limit their use in practice until technology improves in the future. For the function of navigation, the range is relatively short and there are already LPI radars in operation such as the PILOT which makes use of waveform coding. However, no matter which LPI technique is used, the introduction of radar cross section reduction techniques will make LPI operation less effective.

The general system sensitivity requirement for the detection of current and projected LPI radars is found to be on the order of – 100 dBmi which cannot be met by current EW receivers. However with some modification to current narrow–band EW channelizers in terms of reduced video bandwidth, the sensitivity can be improved for LPI radar detection.

Three general LPI ESM architectures, using narrow–band channelizers, superhet and acousto–optic receivers, have been examined in this report for shipborne applications. They have shown some promise in terms of providing the sensitivity and capability in an environment where both conventional and LPI signals are present.

TABLE OF CONTENTS

TABLE OF CONTENDS (cont.)

LIST OF FIGURES

LIST OF FIGURES (cont.)

LIST OF TABLES

1.0 INTRODUCTION

Most current radars are designed to transmit short duration pulses with relatively high peak power. These radars can be detected easily by the use of relatively modest EW intercept receivers which are specifically designed for the interception of this type of radar signals. The intercept of radar transmissions ultimately leads to vulnerability through the use of either antiradiation missiles or ECM. However by using LPI techniques, it is possible to design a low probability of intercept (LPI) radar which is effective against current intercept EW receivers.

There are a number of LPI techniques a radar can employ[1]. These may include low sidelobe antennas, infrequent scanning, power control when tracking a closing target (as range is reduced, the radar power is also reduced), making use of waveform coding to provide transmitting duty cycles approaching one (to reduce peak power while maintaining the required average power) and using frequency hopping to force the interceptor to consider more of the spectrum in attempting to characterize the radar.

In this report, an analysis is presented on current and projected LPI radar signals and the detection of these signals by EW receivers. The analysis starts with an introduction on the difference in detection between a radar receiver and an EW receiver. The ratio of the radar detection range to the EW intercept receiver detection range is then derived. The detection range of three typical radars for the functions of (a) search, (b) anti–ship missile (ASM) seeker and (c) navigation, is then examined against the detection range of three typical EW receivers. The purpose is to evaluate the effectiveness of potential LPI techniques, such as waveform coding, antenna profile control and power management, that a LPI radar may employ against conventional EW receivers. The results are then used to define the requirements for EW receiver on the detection of current and projected LPI signals. Finally some ESM receiver architectures are considered for the detection of these LPI signals.

The analysis to be presented in this report is by no means rigorous. Due to the complexity of specific applications that a radar designer has to address, only simple analysis is used to illustrate the key features considered. This report uses examples analyzed mainly from the detection point of view and other constraints that a radar designer may have to face are not examined. In addition, the numbers used throughout the report are only approximate figures due to the assumptions made and in most cases only free–space propagation is considered.

2.0 LPI RADAR VERSUS INTERCEPT RECEIVER

In this section, the ratio of the intercept receiver detection range to the radar detection range is derived. This ratio is then analyzed as a function of the radar antenna pattern (mainlobe intercept versus sidelobe illumination), radar signal waveform and the type of EW intercept receiver.

2.1 Intercept Receiver Detection Range Versus Radar Detection Range

Using subscript r to denote the LPI radar, the maximum free–space radar range (R_r) of a radar is given by [2]

$$R_r = \left\{ \left[P_{Tr} L_r G_{Tr} G_{Rr} \lambda^2 \sigma \right] / \left[(4\pi)^3 S_{min,r} \right] \right\}^{1/4} \tag{1}$$

where P_{Tr} is the output power from the radar transmitter, L_r is the transmission line loss between the transmitter tube and the antenna terminal and is ≤ 1, G_{Tr} is the power gain of the transmitting antenna, G_{Rr} is the antenna gain of the receiving antenna, σ is the target radar cross section, λ is the wavelength, and $S_{min,r}$ is the minimum detectable signal which is related to the processing gain of the radar receiver.

The optimum filter for detection of a signal in white noise is the matched filter, which has a frequency response equal to the complex conjugate of the radar signal spectrum. With the receiving filter matched to the radar signal spectrum, the output peak signal–to–noise ratio for a single pulse received is given by[3]

$$(S/N)_{mf} = S\,\tau/(\eta F_r) \tag{2}$$

where τ is the pulse duration, F_r is the total noise figure of the radar system which includes the loss between the receiving antenna terminal and the receiver, η is the noise spectral density at the input of the receiver and S is the input received signal power. Many pulses are usually returned from any particular target on each radar scan and thus can be used to improve detection. For n equal pulses integrated incoherently, the minimum detectable signal power becomes

$$S_{min,r} = (\eta F_r)/\tau\,(S/N)_{mf,n} = KT_r F_r/\tau\,(S/N)_{mf,n} \tag{3}$$

where K is the Boltzmann's constant, $(S/N)_{mf,n}$ is the signal–to–noise ratio of one of the n equal pulses that are integrated to produce the required probability of detection for a specified probability of false alarm, and T_r is the radar receiver noise temperature. Substituting Eq.(3) into Eq.(1) yields

$$R_r = \left\{ \left[P_{Tr} \tau\, L_r G_{Tr} G_{Rr} \lambda^2 \sigma \right] / \left[(4\pi)^3 F_r K T_r\, (S/N)_{mf,n} \right] \right\}^{1/4} \tag{4}$$

Equation (4) simply states that the maximum radar range is directly proportional to the fourth root of the energy $(P_{Tr}\tau)$ transmitted.

For a EW radar intercept receiver, a detection decision is made on the basis of a single pulse. Using the subscript i to denote the interceptor, the maximum detection range is

$$R_i = \left\{ \left[P_{Tr} L_i G_{Ti} G_i \lambda^2 \right] / \left[(4\pi)^2 (S/N)_i \, N \right] \right\}^{1/2} \tag{5}$$

where G_{Ti} is the antenna gain of the radar antenna in the direction of the radar interceptor, G_i is the antenna gain of the interceptor, L_i is the loss between the receiving antenna terminal and the input of the receiver, $(S/N)_i$ is the signal–to–noise ratio needed to produce the required probability of detection for a specified probability of false alarm, and N is the effective input noise power given by

$$N = K \, T_i \, B_i \, F_i \tag{6}$$

where F_i is the receiver noise figure, T_i is the receiver temperature (taken to be at room temperature 290° K) and B_i is the equivalent noise bandwidth of the intercept receiver. The equivalent noise bandwidth is going to be examined in details in Section 2.2. The output signal–to–noise ratio $(S/N)_i$ can be calculated by assuming that the minimum detectable signal power competes with the effective input noise power of magnitude

$$S_{min,i} = (S/N)_i \, (K T_i B_i F_i) \tag{7}$$

In Eq.(7), the minimum detectable signal power has been assumed to be linearly proportional to the required output signal–to–noise ratio. It is true when the detection process is linear and the total output noise is dominated by the input noise. However as will be pointed out in Section 2.2 , Eq.(7) is only approximately true for square law detection. Substituting Eq.(7) into Eq.(5) yields

$$R_i = \left\{ \left[P_{Tr} L_i \, G_{Ti} \, G_i \lambda^2 \right] / \left[(4\pi)^2 (S/N)_i \, (K T_i B_i F_i) \right] \right\}^{1/2} \tag{8}$$

A difference between Eq.(4) and Eq.(8) is that for the intercept receiver, the maximum detection range is directly proportional to the square root of the peak power (P_{Tr}) transmitted and not the energy. Dividing Eq.(8) by Eq.(4) , the ratio of the intercept receiver detection range to the radar detection range is

$$R_i/R_r = \left\{ 1/(4\pi) \left[P_{Tr} \tau/(K T_i) \right] \left[F_r/F_i^2 \right] \left[L_i^2/L_r \right] \left[T_r/T_i \right] \left[1/(\tau B_i)^2 \right] \left[\lambda^2/\sigma \right] \right.$$
$$\left. \left[(S/N)_{mf,n}/(S/N)_i^2 \right] \left[G^2_{Ti} G^2_i/(G_{Tr} G_{Rr}) \right] \right\}^{1/4} \tag{9}$$

When the same radar antenna is used for both transmit and receive, G_{Tr} is approximately equal to G_{Rr} and for an omnidirectional intercept antenna, $G_i = 1$. For a certain energy ($P_{Tr}\tau$) or average power transmitted, Eq.(9) can be simplified and expressed directly in terms of the radar waveform, antenna pattern and radar cross section as

$$R_i/R_r = K_1 \left\{ \left[G_{Ti}/G_{Tr} \right] \left[1/(\tau B_i) \right] \right\}^{1/2} \left[1/\sigma \right]^{1/4} \tag{10}$$

From Eq.(10), the ratio R_i/R_r is directly proportional to the square root of the antenna gain of the radar antenna in the direction of the radar interceptor and inversely proportional to the time–bandwidth factor (τB_i). It is also inversely proportional to the fourth–root of the radar cross section.

Equation (9) can also be expressed directly in terms of the radar maximum detection range by making use of Eqs.(4) and (5) to give

$$R_i/R_r = R_r \left\{ \left[L_i G_{Ti} G_i/(L_r G_{Tr} G_{Rr}) \right] (4\pi/\sigma) \left[F_r T_r (S/N)_{mf,n} \right] / \left[(S/N)_i (F_i T_i B_i \tau) \right] \right\}^{1/2} \tag{11}$$

A quiet radar can be defined as one for which $R_i/R_r \leq 1$. From Eq.(11), for a given set of radar and interceptor parameters, the condition $R_i/R_r \leq 1$ can usually be met at close ranges where R_r is small. In other words, LPI operation is most easily achieved at close ranges. This is simply due to the fact that the radar range is proportional to the fourth root of the power while the interceptor range is proportional to the square root of the power. For the case $R_i/R_r = 1$, the radar cannot be intercepted beyond the range at which it can detect targets itself. For this case, the radar range is given by

$$R_r = \left\{ \left[L_r G_{Tr} G_{Rr}/(L_i G_{Ti} G_i) \right] \left[\sigma/(4\pi) \right] \left[(S/N)_i \, (F_i T_i \tau B_i) \right] / \left[F_r T_r \, (S/N)_{mf,n} \right] \right\}^{1/2} \tag{12}$$

Expressing directly in terms of the signal waveform, antenna pattern and radar cross section, and with $G_{Tr} = G_{Rr}$ and $G_i = 1$, Eq.(12) can be simplified to

$$R_r = K_2 \left\{ \left[G_{Tr}^2/(G_{Ti}) \right] \sigma \, (\tau B_i) \right\}^{1/2} \tag{13}$$

In the case where the interceptor is located in the sidelobes of the LPI radar, the LPI radar range is directly proportional to the gain of the antenna when the sidelobe levels are assumed to be of 0 dBi. The LPI radar range can be increased by employing an antenna with a higher gain , however the tradeoffs are larger antenna size and a longer time to search the same volume. For the tracking case, $G_{Tr} = G_{Ti}$, the LPI radar range is only proportional to the square root of the antenna gain. In Eq.(13), the LPI radar range is also directly proportional to the square root of the time–bandwidth factor and the radar cross section.

2.2 Time–bandwidth Factor

In Section 2.1, both the ratio of R_i/R_r and the radar LPI range have been expressed directly in terms of the time–bandwidth factor (τB_i). It is important to keep in mind that B_i is the equivalent noise bandwidth of the intercept receiver while τ is the pulse duration of the radar signal. In this section, the time–bandwidth factor is examined in details in terms of the type of EW intercept receiver and radar signal waveform.

Dividing Eq.(7) by Eq.(3), the ratio of the minimum detectable signal level of the intercept receiver to the minimum detectable signal level of the radar is

$$S_{min,i}/S_{min,r} = (\tau B_i) \left[(S/N)_i/(S/N)_{mf,n}\right] \left[(F_i T_i)/(F_r T_r)\right] \tag{14}$$

As can be seen from Eq.(14), the processing gain of the radar receiver over the intercept receiver is directly proportional to the time–bandwidth factor (τB_i).

For a crystal video receiver using a square law detector, the tangential sensitivity of the crystal video receiver is [4,5]

$$\begin{aligned} TSS &= KT_i F_i \left\{ 6.31\, B_v + 2.5 \left[2\, B_{RF}\, B_v - B_v^2 + AB_v/(GF_i)^2 \right]^{1/2} \right\} \\ &= K\, T_i\, F_i\, B_{TSS} \end{aligned} \tag{15}$$

where $B_{RF} \geq 2\, B_v$, B_{RF} is the RF bandwidth, B_v is the video bandwidth and B_{TSS} is the tangential equivalent noise bandwidth so that a zero dB input signal–to–noise ratio will produce a tangential sensitivity (approximately 8 dB S/N ratio) at the output of the detector. It must be noted that the detection process is not linear and an increase in the input signal does not proportionally improve the output signal–to–noise ratio. The output noise will increase when the input signal increases, because the output noise contains a signal–noise cross product term [5]. A is the diode parameter[4], and G is the gain and F_i is the noise figure of the amplifiers ahead of the detector respectively.

When a receiver is input noise limited, the last term $AB_v/(GF_i)^2$ in Eq.(15) is negligible. For this case, the tangential sensitivity is plotted in Figs.1 and 2 as a function of B_{RF} and B_v with $F_i = 0$ dB. As can be seen from the plots, the sensitivity can be greatly improved by appropriately reducing both the RF and video bandwidths.

2.2.1 Crystal Video Detector Receiver With Preamplifier

For $B_{RF} >> B_v$ and large G so that the receiver is input noise limited, the tangential noise equivalent bandwidth is

$$B_{TSS} \approx 2.5 \left[2\, B_{RF}\, B_v\right]^{1/2} \tag{16}$$

and the equivalent noise bandwidth as defined by Klipper [6] is

$$B_e \approx \left[2\, B_{RF}\, B_v\right]^{1/2} \tag{17}$$

With the above definition of equivalent noise bandwidth which is 2.5 times smaller than B_{TSS}, a 4–dB input signal–to–noise ratio is required to produce the same tangential sensitivity at the output of the receiver.

In this analysis, the equivalent noise bandwidth (B_i) is taken to be

$$B_i \approx 0.4\, B_e = 0.16\, B_{TSS} = 0.4 \left[2\, B_{RF}\, B_v\right]^{1/2} \tag{18}$$

With this equivalent noise bandwidth, an 8–dB input signal–to–noise ratio is required to produce a tangential sensitivity at the output of the detector. Due to the non–linear detection process, the minimum detectable signal power is no longer linearly proportional to the required output signal–to–noise ratio as given by Eq.(7). In other words, a 0–dB input signal–to–noise ratio at the same equivalent bandwidth does not produce a 0–dB signal–to–noise ratio at the output. In order to relate exactly the input to output signal–to–noise ratios, the signal–noise cross product term [5] has to be taken into account.

In an intercept receiver, B_v is designed for the shortest anticipated pulse width (τ_{min}). Substituting $2B_v = 1/\tau_{min}$ into Eq.(18) ,the processing gain of the radar receiver over the intercept receiver (τB_i) is

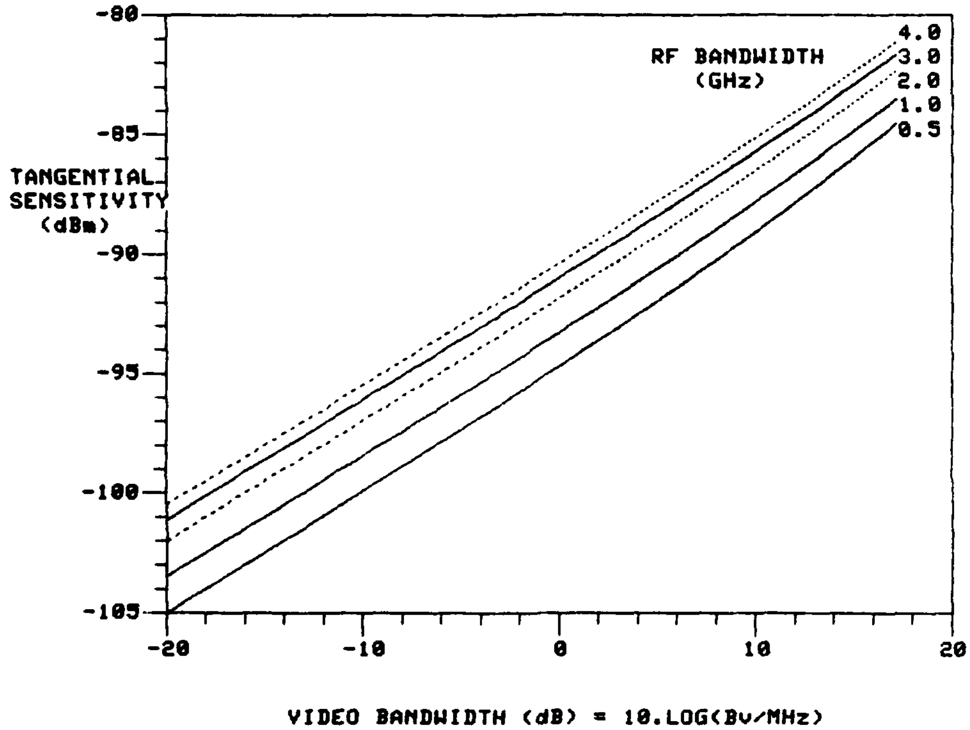

FIGURE 1. TANGENTIAL SENSITIVITY AS A FUNCTION OF RF AND VIDEO BANDWIDTHS ($F_i = 0$ dB)

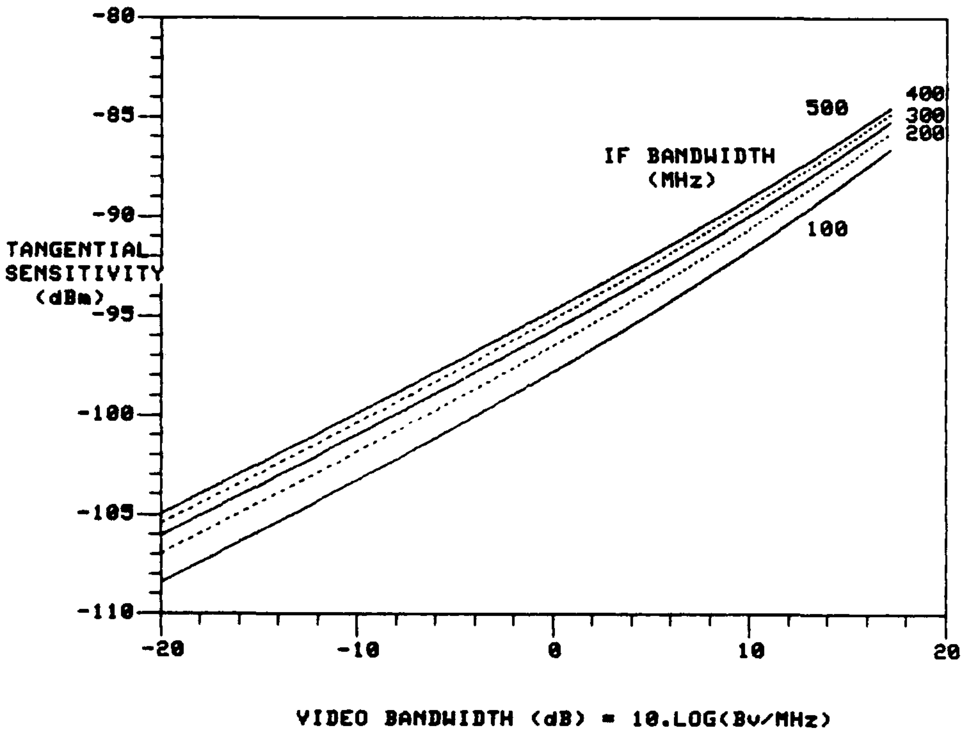

FIGURE 2. TANGENTIAL SENSITIVITY AS A FUNCTION OF IF AND VIDEO BANDWIDTHS ($F_i = 0$ dB)

$$B_i\tau = 0.4 \left[2\, B_{RF}\, B_v\right]^{1/2} \tau$$

$$= 0.4\, M \left[B_{RF}\tau_{min}\right]^{1/2} \qquad (19)$$

where $M = \tau/\tau_{min}$ [7] is the pulse width mismatch factor. The processing gain can be increased by simply using a radar waveform with a longer duration. However as given in Eq.(9) the same average power has to be used. This is done by reducing the peak power so that the average transmitted power will remain constant. In other words, LPI operation of a radar can be achieved by increasing the duration of the pulse width and at the same time by reducing the peak power so that the same average power is transmitted. If the duration of a simple pulse is increased, the resultant radar range resolution will be reduced (range resolution = $c\tau/2$, where c is the speed of light). As a result, in order to maintain the same range resolution, a wideband signal has to be used so that the return signal can be compressed [3]. Substituting $\tau = \beta\, \tau_{eff}$ into Eq.(19), yields

$$B_i\tau = 0.4\, \beta\, M_p \left[B_{RF}\tau_{min}\right]^{1/2} \qquad (20)$$

where $M_p = \tau_{eff}/\tau_{min}$ is the mismatch factor relating the compressed pulse width (τ_{eff}) to the minimum pulse width (τ_{min}) anticipated by the intercept receiver, and β is the pulse compression ratio or time–bandwidth product of the waveform. This time–bandwidth product is not to be confused with the time–bandwidth factor which has been used extensively up to now.

When the same range resolution is required, M_p must be equal to M . In order to maintain the same average transmitted power when the uncompressed pulse width has been increased by β, the peak power has to be reduced by the same factor. As a result, the processing gain of the radar receiver over the intercept receiver is increased by the factor β and the ratio R_i/R_r [Eq.(9)] is also reduced by the square root of β.

It has been shown that a radar can be made more effective in terms of LPI operation, by using a pulse compression waveform where the duration of the transmitted signal is increased while the range resolution is maintained. However the use of a longer pulse compression signal as compared to a simple pulse has some disadvantages. One of the main drawbacks is that the long uncompressed pulse can restrict the minimum range and the ability to detect close–in targets [3]. This disadvantage will be considered in greater detail in Section 3.

When B_{RF} is larger than the bandwidth of the signal, more noise will enter the intercept receiver and the time–bandwidth factor will also increase. If B_{RF} is smaller than the bandwidth of the signal, less noise will be detected by the intercept receiver. On the other hand, the amount of signal power detected will also decrease and the signal will also be distorted. As B_{RF} is reduced further, the noise generated by the detector will eventually dominate and the noise can not be reduced further. The best compromise in terms of maximizing the output signal–to–noise ratio and to retain the signal fidelity is to make B_{RF} approximately equal to the bandwidth of the signal.

2.2.2 Crystal Video Detector Receiver Without Preamplifier

If there is no preamplifier ahead of the detector, the noise equivalent bandwidth is

$$B_i \approx 0.4/(GF_i)\left[AB_v\right]^{1/2} \tag{21}$$

where A is the diode parameter typically of the order of 10^{14} MHz [4] and

$$\begin{aligned}\tau B_i &= 0.4/(GF_i)\left[A\tau^2/(2\tau_{min})\right]^{1/2}\\ &= 0.4\ M/(GF_i)\left[A\tau_{min}/2\right]^{1/2}\end{aligned} \tag{22}$$

In terms of pulse compression waveform

$$\tau B_i = 0.4\ \beta\ M_p/(GF_i)\left[A\tau_{min}/2\right]^{1/2} \tag{23}$$

2.2.3 Channelized Receiver

In a channelized receiver such as a filter bank, the channel bandwidth B_{RF} is not normally larger than $2B_v$. If the receiver is input noise limited, then the equivalent noise bandwidth is

$$B_i \approx B_v + 0.4\left[2\ B_{RF}\ B_v - B_v^2\ \right]^{1/2} \tag{24}$$

and

$$\tau B_i = M\left\{1/2 + 0.4\left[\ B_{RF}\tau_{min} - 1/4\right]^{1/2}\right\} \tag{25}$$

In terms of pulse compression waveform where the bandwidth of the signal falls within the channel bandwidth of one of the filters

$$\tau B_i = M_p\ \beta\left\{1/2 + 0.4\left[\ B_{RF}\tau_{min} - 1/4\right]^{1/2}\right\} \tag{26}$$

3.0 EXAMPLES OF LPI RADAR VERSUS EW RECEIVERS

The case of a LPI radar versus EW receivers is illustrated by considering three functions of a radar in the areas of (a) search (b) ASM RF seeker and (c) navigation.

The three different types of EW intercept receivers as discussed in Sections 2.2.1 to 2.2.3 are assumed to be operating on a shipborne platform. The characteristics of the EW receivers to be used in the examples are outlined as follows:

(a) IFM Receiver or Crystal Video With Preamplifier

This receiver type is assumed to have the following characteristics:

B_{RF} = 4 GHz
B_v = 10 MHz
L_i = 15 dB
F_i = 7 dB

Using Eq.(18), the equivalent noise bandwidth (B_i) is calculated to be 113.1 MHz.

Approximately a 11–dB input signal–to–noise ratio is needed to produce a Pfa of 10^{-8} and a probability of detection of 95% at the output of the square law detector [4]. In this analysis $(S/N)_i$ = 12 dB is used for the three different types of receivers. The sensitivity of this receiver is

$$\begin{aligned}\text{Sensitivity} &= F_i K T_i B_i (S/N)_i \\ &= -74.5 \text{ dBm}\end{aligned} \tag{27}$$

For an omnidirectional antenna G_i = 0 dBi and with L_i = 15 dB , the system sensitivity is

$$\text{System sensitivity} = F_i K T_i B_i (S/N)_i L_i / G_i = -59.5 \text{ dBmi} \tag{28}$$

The system sensitivity in dBmi is defined as the minimum detectable signal in dBm required at the antenna aperture of the receiving system and referenced to an isotropic antenna gain of 0 dBi.

(b) Crystal Video Receiver Without Preamplifier

The crystal video receiver without preamplification is the simplest microwave receiver. Most current operational radar warning receivers (RWR) are some form of a crystal video receiver without amplification. The antenna gain is assumed to be 10 dBi and the loss between the antenna terminal and the receiver is 2 dB. The RF bandwidth is 4 GHz and the video bandwidth is 10 MHz. Using Eq.(21), the noise equivalent bandwidth

(B_i) is calculated to be 1.26 x 10^7 MHz. Using the same $(S/N)_i$ value as for the case of the IFM receiver, the sensitivity of the receiver at the terminal of the antenna is –29 dBm. When it is referenced to an isotropic antenna, the sensitivity is – 39 dBmi.

(c) Channelized Receiver

The channelized receiver used in this example is assumed to have a channel width of 20 MHz and a video bandwidth of 10 MHz. Using Eq.(24), B_i is computed to be 16.93 MHz. The rest of the parameters are assumed to be the same as for the case of the IFM receiver. Due to a reduction in the noise equivalent bandwidth, the only difference is an improvement in sensitivity of 8.25 dB and the system sensitivity is now – 67.75 dBmi.

3.1 Search Radar

For a modern medium–range search radar, the following are typical parameters:

Peak Power(P_{Tr})	250 kW
Pulse Width (τ)	1 μs
Frequency	9 GHz
Antenna Gain $G_{Tr} = G_{Rr}$	30 dBi (boresight)
Sidelobe Level	30 dB down from boresight
Ultra–low Sidelobe Level	50 dB down from boresight
Noise Figure of Receiver	5 dB
Line Loss (L_r)	5 dB
$(S/N)_{mf,n}$	9.5 dB for n = 5 pulses integrated, with Pfa of 10^{-10} and a probability of detection of 95%, Swerling Case 1

Substituting the above parameters into Eq.(4), the maximum free–space range for detecting a target with a radar cross section (σ) of 1 m^2 is calculated to be 18.75 km. The ratio of the detection range of an interceptor to the radar detection range is then calculated by using Eq.(9) for the three different types of EW receivers.

It is to be noted that the R_i/R_r ratio is computed by assuming free–space propagation. When propagation effects, such as multi–path and attenuation due to rain, are taken into account the radar will be affected more than the interceptor due to two–way propagation. In addition, the signal–to–noise ratio $(S/N)_{mf,n}$ as specified above is needed merely for the purpose of detecting the target only. If finer details on the target such as angular information is to be extracted, then a higher signal–to–noise ratio is required. There are also other factors that the radar designer may have to consider such as for ECCM purposes where a higher signal return is needed. Moreover, the value of the measured radar cross section is a strong function of aspect angle and is also subjected to glint and noise. As a result, the required output signal–to–noise ratio for the radar can be much higher. On the other hand, if the intercept receiver is required to do detailed analysis on the radar signal such as angular information and intrapulse modulation, a higher output signal–to–noise ratio is also required. Another important factor to consider is that from Eq.(9), the R_i/R_r ratio is directly proportional to the fourth root of the signal–to–noise ratio at the radar receiver while it is inversely proportional to the square root of the signal–to–noise ratio at the intercept receiver. In general, the output signal–to–noise ratio required by the radar receiver may be higher than that of the intercept receiver, the net effect on the ratio of R_i/R_r can be small. In the following analysis , the ratio of R_i/R_r is calculated by using the free–space propagation model and used as a baseline for comparison.

For the IFM receiver, when the intercept receiver is illuminated by the boresight of the radar antenna, the ratio of R_i/R_r is computed to be 20.21 dB. If the sidelobe of the search radar is intercepted, R_i/R_r is reduced by 15 dB which is the square root of the sidelobe level.

Instead of using a simple pulse waveform, the search radar could use a longer pulse compression waveform to achieve the same range resolution and be able to reduce the peak power transmitted as discussed in Section 2.2.1. Substituting Eq.(20) into Eq.(9), R_i/R_r is found to be inversely proportional to the square root of the time–bandwidth product (β). It is this relationship which is plotted in Fig. 3 as a function of mainbeam and sidelobe intercepts.

For the crystal video receiver without preamplification, the ratio of R_i/R_r is found to be 10.0 dB when the intercept receiver is illuminated by the boresight of the radar antenna. For the channelized receiver, the ratio of R_i/R_r is 24.3 dB.

The ratio of R_i/R_r is plotted in Fig. 3 for the three different receiver types. It is plotted as a function of the time–bandwidth product and the antenna sidelobe levels for a target with a radar cross section of 1 square meter. It is noted that for each type of receiver, the time–bandwidth product must meet the constraints outlined in Sections 2.2 to 2.2.3 for both B_{RF} and τ. For example in the case of the crystal video receiver, the pulse width (τ) can not be less than 0.05 μs and the total RF bandwidth has to be less than 4 GHz.

From the plot, the medium–range search radar is detected by all of the receivers when the mainbeam of the radar is intercepted. For a signal with a modest time–bandwidth product of 100, the radar is shown to be effective only against the crystal video receiver without preamplification. For the other two types of receivers, the radar is detected even if a signal with a large time–bandwidth product is used. As a result, it is very difficult to make the radar LPI in operation by using waveform coding alone when the interceptor is illuminated by the mainbeam of the radar. However, with the combination of

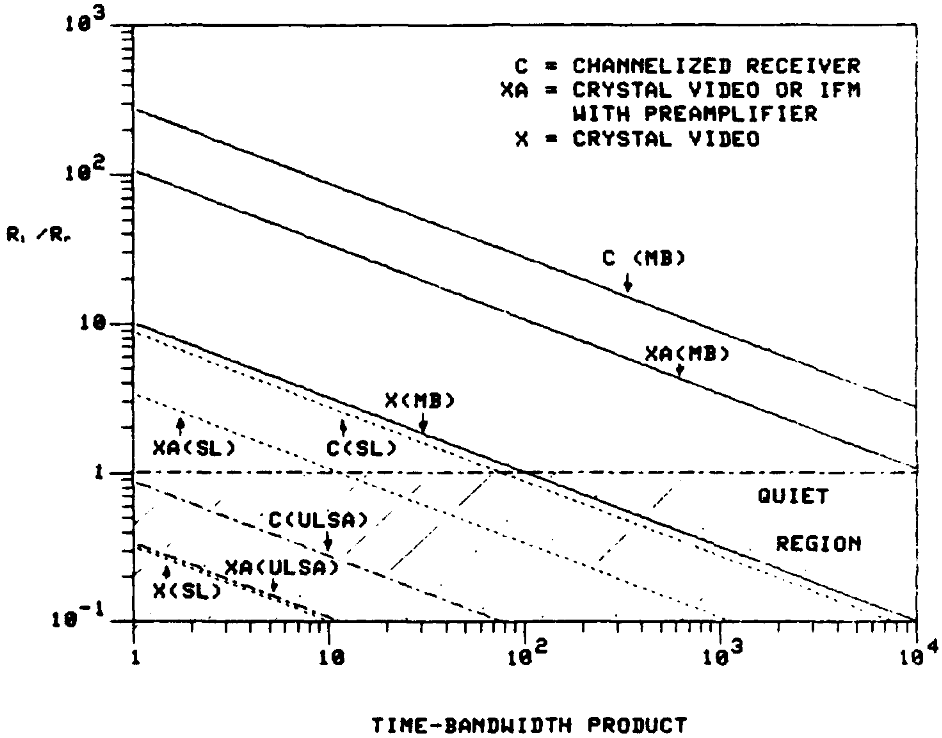

FIGURE 3. RATIO OF INTERCEPT RECEIVER RANGE TO RADAR RANGE FOR WIDE–BAND WAVEFORMS OF A SEARCH RADAR

antenna beamshape control and waveform coding, the medium–range search radar can avoid being detected by the three different types of receivers when the interceptor is located in the sidelobes. This may not be a useful situation as sidelobe interception is not strictly necessary to carry out the interception process. On the other hand, this situation is still quite useful against anti–radiation missiles and ECM (jamming) which are directed towards the radar antenna through its sidelobes.

The general conclusion which can be drawn from this example is that it is very difficult for the search radar to operate effectively in a LPI mode of operation. This is mainly due to the stringent requirement that the radar has to search for a target with a small radar cross–section and over a relatively long range.

The employment of a large time–bandwidth product waveform for LPI applications enables the radar to transmit a much lower peak power signal and the same range resolution is obtained by using pulse compression techniques. This implies the use of longer duration and thus higher duty–cycle signals. The use of a longer duration signal requires that good isolation is needed between the radar transmitter and radar receiver when the radar is transmitting. This additional requirement must be met in order for this type of waveform to be employed. In general, a better isolation is achieved if two separate antenna apertures are used for simultaneous transmit and receive.

There are cases where a high range resolution is required such as for the detection of periscope and snorkel classes of targets. A range resolution of the order of 1.5 feet is usually needed and which closely matches the physical dimensions of these targets. If a linear FM pulse compression signal is used, a range resolution of 1.5 feet will require a compressed pulse width of 3.28 ns and a linear total frequency deviation of 300 MHz. For most targets, a range resolution corresponding to around 0.1 μs is adequate and thus the total frequency deviation is expected to be about 10 MHz. As a result, the projected LPI signal waveforms are of longer durations and thus of higher duty cycles. The use of a pulse doppler can also provide some coherent processing gain from pulse to pulse, but the range ambiguity will restrict its maximum range of operation.

3.2 ASM RF Seeker

3.2.1 Typical ASM Seekers

The path loss versus range is plotted in Fig. 4 for a typical medium surface–to–surface missile (SSM) ASM seeker. The peak power of the transmitter (P_{Tr}) is 90 kW and the antenna gain (G_{Tr}) is 30 dBi. The radar seeker is usually activated at a range of approximately 30 km and the lock–on range is approximately 20 km. A radar ESM system sensitivity threshold is also plotted on the same figure and is assumed to be of an IFM receiver type with a sensitivity of – 59.5 dBmi. The free–space path loss and the loss including propagation effects are plotted. The location and depth of nulls are determined by factors such as the polarization and frequency of the electromagnetic wave, the reflectivity and roughness of the sea surface, and also on the height of the radar antenna (H_r) and height (H_i) of the receiving antenna of the interceptor. As can be seen

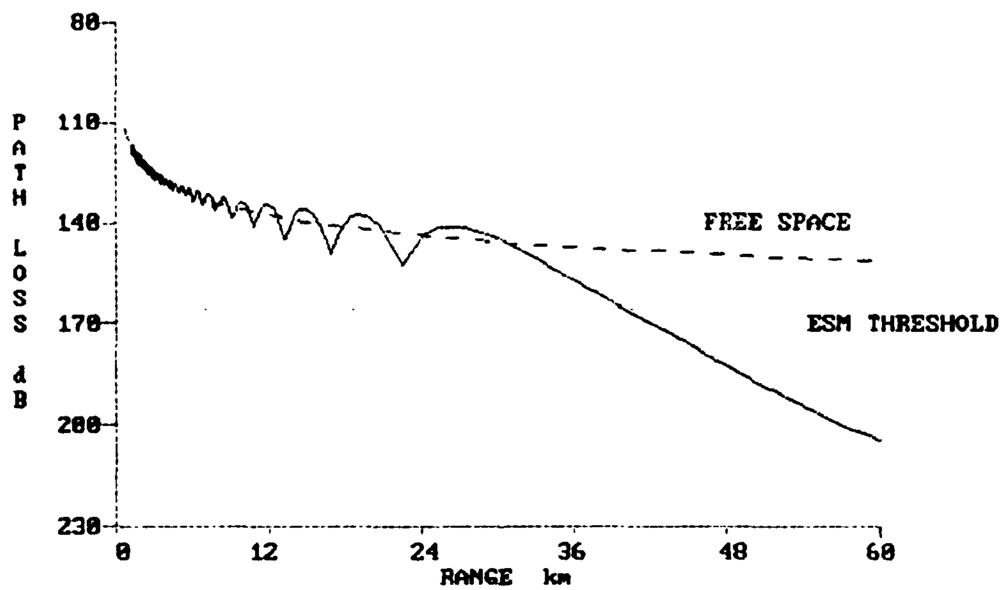

FIGURE 4. PATH LOSS VERSUS RANGE FOR A TYPICAL MEDIUM SSM ASM SEEKER [H_r = 20 m, H_i = 25 m, Peak Power = 90 kW, G_t = 30 dBi, Wind = 20 kts, ESM threshold (G_r = 0 dBi, Cable Loss = 15 dB, Receiver TSS Sensitivity = – 74.5 dBm)]

from the plot, the radar ESM receiver can detect the radar seeker with at least a 20 dB margin above the threshold setting.

Figure 5 shows another plot for a more modern medium SSM ASM seeker. The main difference between this case and the previous one is that this missile transmits a lower peak power and is flying at a lower altitude with an activation range of only 10 km. Comparing to Fig. 4 , the received power level is at least 10 dB lower.

Figure 6 shows the free–space received power level versus range for a number of typical ASM seekers deployed. The general conclusion, is that these seekers can be detected easily using both the IFM and channelized receivers. For the crystal video receiver without preamplifier , it may have some difficulty against the more modern short range seekers with lower peak powers.

3.2.2 Modern Seeker

The following is a list of typical parameters of a modern ASM seeker [8]

Peak Power (P_{Tr})	25 kW
Antenna Gain (G_{Tr})	30 dB
Frequency	10 GHz
Beamwidth	6.0 degrees
PRF	4000 Hz
Number of Pulses on Target	100
Range Resolution	15 m
Pulse width (τ)	100 ns
Noise Figure of receiver	5 dB
Line Loss (L_r)	3 dB

With 100 pulses integrated, $(S/N)_{mf,n}$ must be 1 in order to produce a detection probability of 95% and Pfa of 10^{-6} [8]. When this radar is used against the IFM receiver, R_i/R_r is calculated to be 3.64 or 5.6 dB for a target with a radar cross section of 10,000 m^2. The R_i/R_r ratio for mainbeam intercept is plotted in Fig. 7 against the three different types of EW receivers. From the plot, it is seen that for such a large target cross section, the radar can operate in the quiet region against the IFM and channelized receivers with a waveform of a modest time–bandwidth product. The crystal video receiver without preamplification cannot detect the radar at all.

The free–space radar detection range as calculated by using Eq.(4) is 145 km. The actual activation range of the radar for ASM applications is usually much shorter and typically from 15 to 30 km.

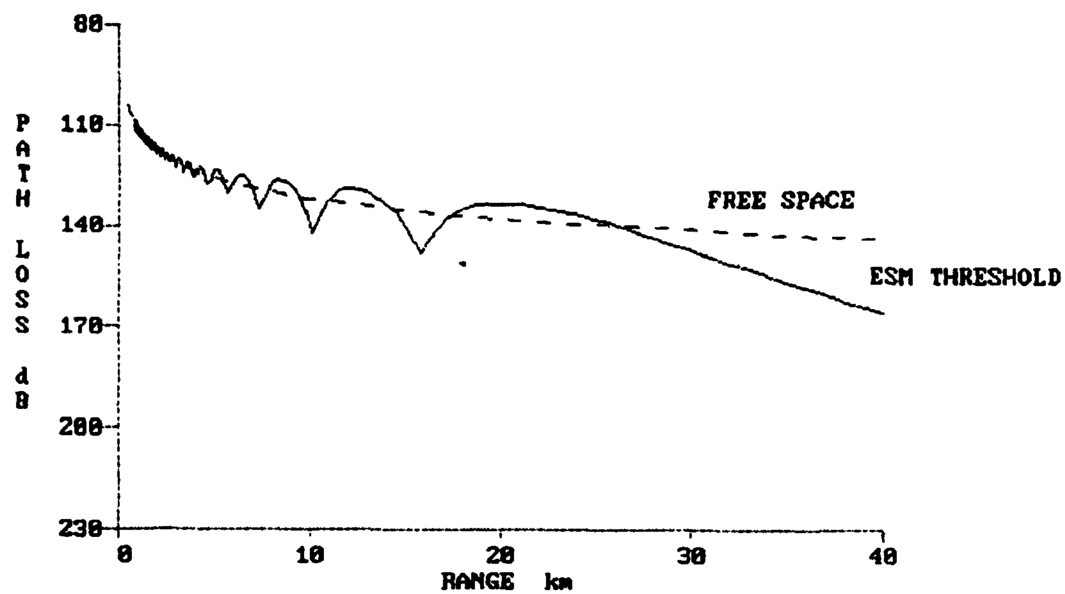

FIGURE 5. PATH LOSS VERSUS RANGE FOR A TYPICAL MEDIUM SSM ASM SEEKER [($H_r = 10$ m, $H_i = 25$ m, Peak Power = 30 kW, $G_t = 20$ dBi, Wind = 20 kts, ESM threshold ($G_r = 0$ dBi, Cable Loss = 15 dB, Receiver Sensitivity = -74.5 dBm)]

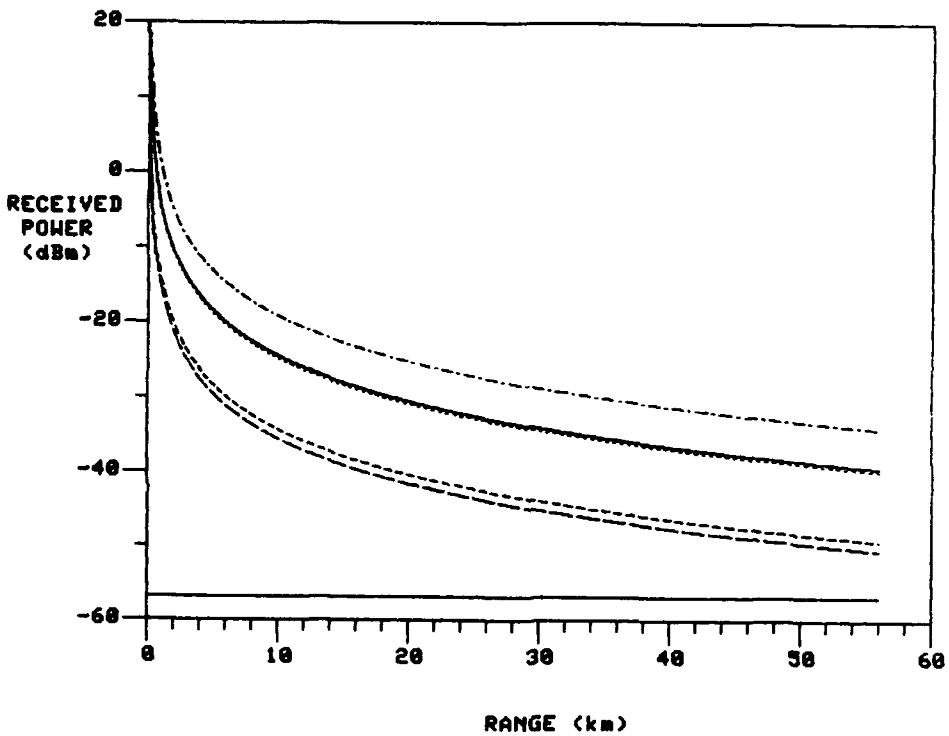

FIGURE 6. INTERCEPTED POWER LEVEL BY A TYPICAL RADAR ESM RECEIVER VERSUS RANGE FOR SOME ASM SEEKERS

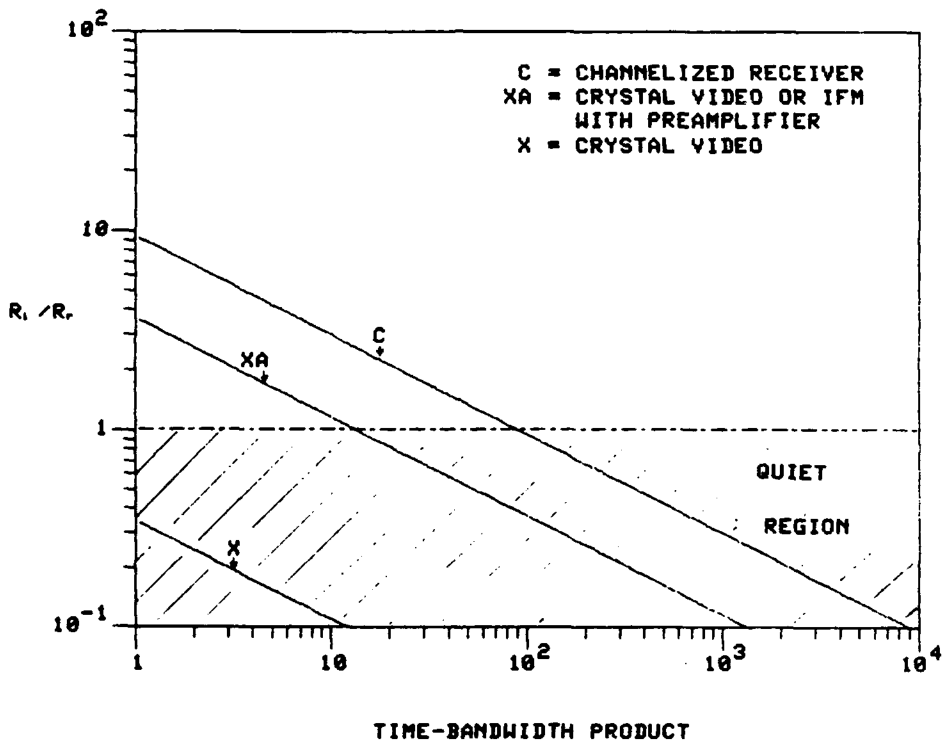

FIGURE 7. RATIO OF INTERCEPT RECEIVER RANGE TO RADAR RANGE FOR WIDE–BAND WAVEFORMS OF A ASM SEEKER

It is also interesting to analyze the return echo back to the radar receiver as a function of range. Figure 8 plots the typical clutter and receiver noise profiles for a sea–skimming trajectory at 10 m against a ship with a radar cross section of 10,000 square meters. Both clutter and receiver noises are expressed in terms of the equivalent radar cross section, i.e. the radar cross section of a target which at the defined range would create a signal of equal return power. The total noise background which is the sum of the receiver noise and clutter return is also shown. The clutter power increases rapidly with decreasing range. This is due to the $1/R^3$ relationship and the fact that the clutter scattering coefficient rises due to the higher values of the grazing angle as the missile approaches the target [3,8]. It shows that the total noise is typically dominated by the receiver noise at long ranges and the clutter return at short ranges. At a range of about 15 km, the signal–to–noise ratio is about 40 dB which is much more than is required for free–space detection. In a real environment, the actual signal–to–noise ratio is likely to be much less than the 40 dB due to factors mentioned in Section 3.1. At a range of 15 km, the power received at the target is –34 dBm for a system with an antenna gain of 0 dBi. At this received power level, all of the three different types of EW receivers considered in this report can detect the radar. If an additional 30 dB margin is assumed for the return signal, a 10 dB power reduction can be used to reduce the peak transmitted power for LPI operation. When the peak transmitted power is reduced by 10 dB, the power received by the target is – 44 dBm. At this received power level, the crystal video receiver without preàmplification cannot detect the radar. If further reduction in peak power is required, waveform modulation has to be used. If a LPI signal waveform of 30 dB is used so that the uncompressed pulse width is increased from 100 ns to 100 μs, the peak power required will be reduced further from the original value of 25 kW to 2.5 W. The required receiver sensitivity is now – 74 dBmi which are below the detection of all three different types of EW receivers.

Once LPI operation is achieved at the maximum activation range, the transmitted power has to be controlled if LPI operation is to be maintained. This is due to the fact as the missile gets closer to the target, the received power by the intercept is going to increase. As the range between the missile and target is reduced, the received power by the target increases with the square of the range while the return signal to the radar is proportional to the fourth power of range due to two–way propagation. In other words, once LPI operation ($R_i/R_r = 1$) is achieved at the maximum range, then the ratio R_i/R_r can be made even smaller as the missile approaches the target.

The use of waveform modulation requires the employment of signals with a longer pulse duration. A long uncompressed pulse can restrict the minimum range of operation as the missile approaches the target. This minimum range occurs when the leakage of the transmitter signal into the receiver is strong enough to cause degradation to the performance of the radar and the receiver becomes inoperable during the transmission of the radar. As a result, the use of long uncompressed signal waveform has its limitation unless very high isolation is achieved between the transmitter and receiver. This problem is much more severe in a missile head where space is very limited.

So far the analysis has been concentrated on a relatively large target size. With radar cross–section reduction techniques, typical radar cross sections are expected to be

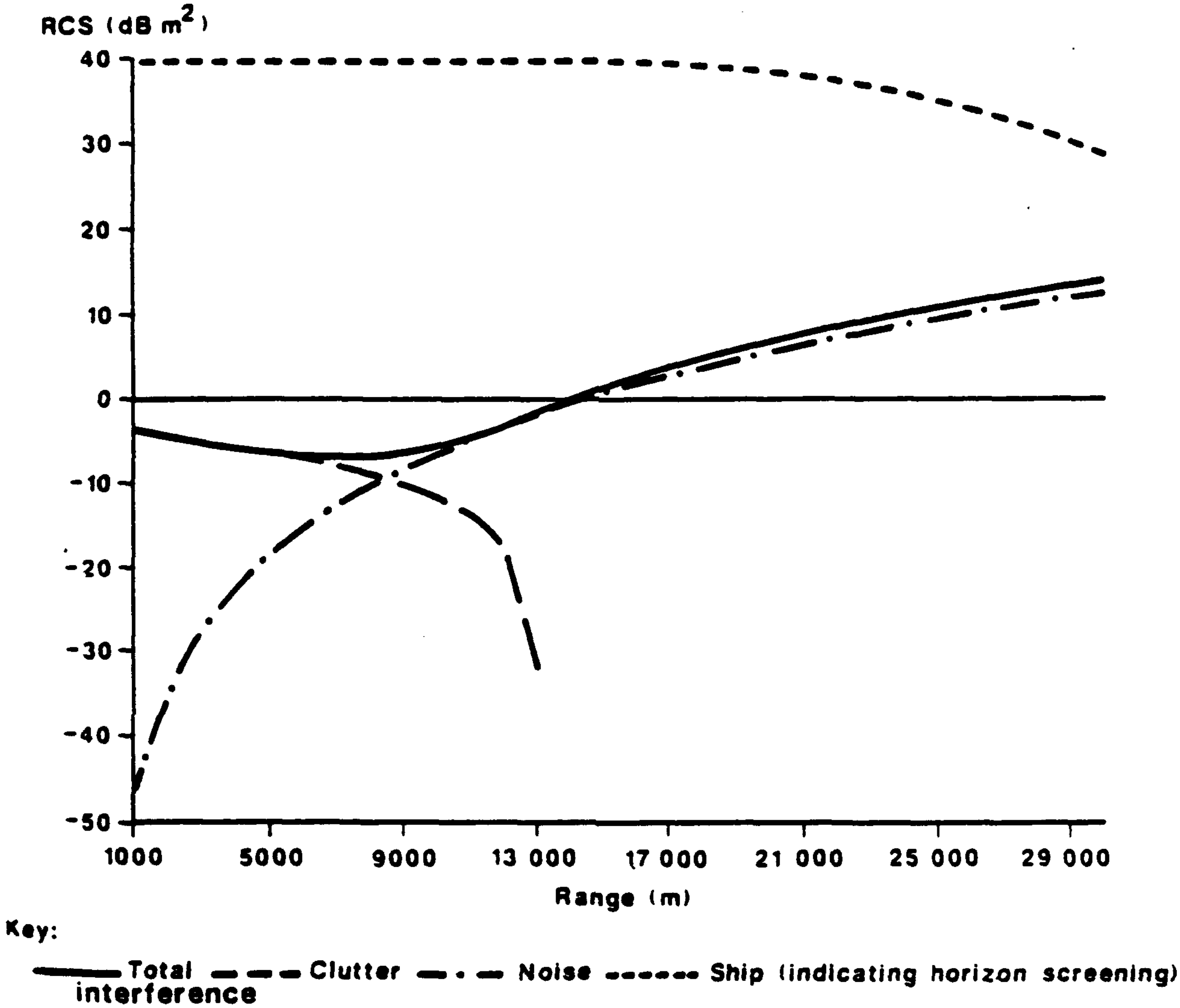

FIGURE 8. INTERFERENCE VERSUS RANGE FOR A SEA–SKIMMING ASM AT 10 m

much smaller. If the target size is greatly reduced, the techniques of both power control and waveform modulation will be less effective.

In current ASM seekers, both power management and waveform coding are not used for a number of practical reasons. The transmitter is usually of magnetron type where there is very little control over the strong output power level and on the waveform modulation. Another reason is due to the complexity and cost associated with the extra processing on the return echo and be able to adjust continuously the output power to tailor each target and the propagation condition in that environment. However these techniques of both power management and waveform coding are increasingly likely to be considered in the future as technology improves.

3.3 Navigation Radar

The received power (assuming an omnidirectional antenna with 0 dBi gain) as a function of free–space range is plotted in Fig. 9 for a convention navigation radar and PILOT radar . Philips in Sweden and Signaal in the Netherlands have developed the "quiet" navigation radars PILOT and SCOUT for covert operations. The peak power of the conventional radar is assumed to be 25 kW. The antenna gains of both conventional and quiet radars are assumed to be 30 dBi and the sidelobes are 30 dB down. PILOT uses a frequency modulated continuous wave (FMCW) transmitter with low peak power (typically 1 Watt). A sweep repetition period of 1 ms is employed and an FFT processor is used to integrate the received signal coherently. Range resolution is achieved by sweeping the frequency of transmission, and processing by the 1024 point FFT to yield 512 range cells per range sweep. In terms of processing gain , PILOT has an equivalent time–bandwidth product of 1024.

The system sensitivity of the IFM is also plotted as a horizontal bar in Fig.9. As can be seen from the plot, the conventional navigation radar can be detected by the IFM receiver for both mainbeam and sidelobe illumination. However for the PILOT radar, the detection range is only 2 km for mainbeam intercept. The system sensitivity requirement for detecting the PILOT radar is approximately − 85 dBmi for mainbeam intercept and − 115 dBmi for sidelobe illumination at a free–space range of 35 km.

4.0 RECEIVER REQUIREMENTS FOR THE DETECTION OF LPI RADARS

Specific examples have been given in Section 3 on the three functions of a radar and some potential techniques in making a radar LPI in operation. In this section, a general discussion is given on the possible LPI signal characteristics and on the projected sensitivity and dynamic range requirements of EW receivers for intercepting these LPI radars. For shipborne EW systems, more sensitive types of ESM receiver such as the IFM and channelized receivers are expected to be used.

It must be emphasized again that the numbers arrived at from the calculations are only approximation subjected to a large margin of variation. It is very difficult to obtain accurate figures because most of the calculations are based on free–space propagation. Even

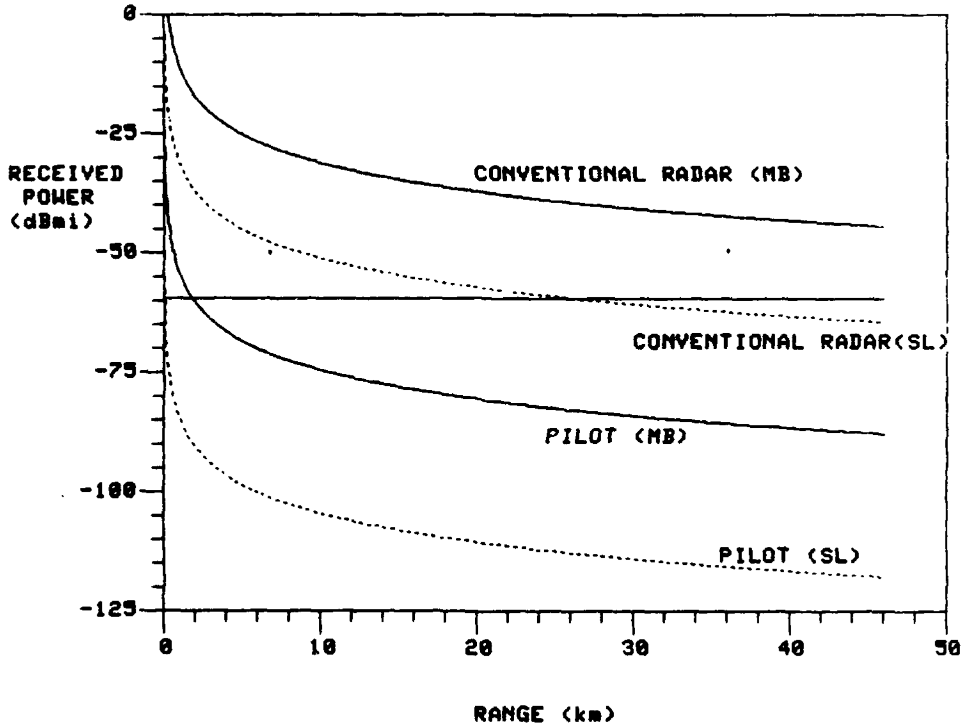

FIGURE 9. INTERCEPTED POWER LEVEL BY A TYPICAL RADAR ESM RECEIVER VERSUS RANGE FOR A CONVENTIONAL NAVIGATIONAL RADAR AND PILOT RADAR

when propagation effects are taken into account, the calculations are very sensitive to the assumed parameters and operating conditions. In addition, the radar problem has only been considered from a very general functional point of view and there may be other constraints imposed by the specific requirement of the radar and its operating environment. As a consequence, judgement should be taken in interpreting the numbers derived in this report.

In a practical situation as illustrated in the search and ASM seeker functions, the radar designer is mainly concerned with meeting the primary function of the radar by maximizing its performance. The task of making the radar LPI in operation against EW receivers has not been the prime motivation factor. In addition, as illustrated by examples on the search and ASM seeker functions, it is not easy to make the radar truly LPI in operation due to technology problem, cost and complexity of implementing the different potential LPI techniques. However with the advancement of technology, it is becoming more and more practical to implement some of the schemes. In addition, there is also a growing concern in the radar community on the intercept of radar transmissions which ultimately leads to vulnerability through the use of either antiradiation missiles or ECM.

4.1 LPI Signal Characteristics

From Section 3, it is concluded that a radar is effective against current EW receivers when a low–peak power and long duration signal is used with a large time–bandwidth product. Large bandwidth signals greater than 10 MHz may not be needed unless very high range resolution is required. This implies that signals of relatively narrow bandwidths and high duty cycles are effective for LPI applications. An effective time–bandwidth product of around 1000 is practicable and can be achieved at the present time. With an effective time–bandwidth of 1000, the peak power of modern radars would be lowered by a factor of 1000 or 30 dB.

If the original conventional pulse width is 0.1 μs, using a time–bandwidth product of 1000 requires that the uncompressed pulse width be 100 μs in order to maintain the same range resolution. If the peak power of the conventional pulse is 25 kW, then the LPI signal would be reduced to 25 W. If this reduced peak power is not low enough to avoid degradation to the performance of the radar receiver, then the minimum range could be limited by the transmitted pulse width. The 100 μs pulse width would give a minimum range of 15 km which is not acceptable for most applications. One solution is to use a even larger time–bandwidth product until the peak power is reduced to an acceptable level to allow simultaneous transmit and receive, and then there would be no limits on the minimum range of operation. Another solution is to use a lower time–bandwidth product, so that the minimum range is can be reduced. If a time–bandwidth product of 100 is used, the minimum range will be reduced to 1.5 km and the peak power is 250 W.

4.2 Sensitivity and Dynamic Range Requirement

In the case of a search radar and for mainbeam intercept, it is impractical to make the radar LPI using waveform modulation alone. The combination of low antenna sidelobe and waveform modulation may be effective against current ESM receivers. For the following discussion, the radar is assumed to be using a LPI signal waveform with a time–bandwidth product of 30 dB. The power received by a EW receiver is directly proportional to the peak power transmitted. If a 10 dB lower time–bandwidth product is used, the required sensitivity for the EW receiver will be reduced by the same amount. From Fig. 3 and for mainbeam intercept, the required system sensitivity is – 49 dBmi and both the IFM and channelized ESM receivers will have no difficulty in detecting the radar. For sidelobe intercept, the required system sensitivity is now –79 dBmi while for ultra–low sidelobe antenna, it is – 99 dBmi.

For the case of an ASM seeker, only the mainbeam intercept is of interest. The radar is again assumed to be using a LPI signal waveform with a time–bandwidth product of 30 dB. As discussed in Section 3.2.2, when only power control of 10 dB is used at the activation range of 15 km, a system sensitivity of – 44 dBmi is required. If a combination of power control and LPI waveform are used, then the system sensitivity requirement is – 74 dBmi at the same activation range of 15 km. If power control is still exercised as the missile is closing in on the target, the power received by the target will even be less. At a range of 7.25 km, the system sensitivity will be –80 dBmi. With cross–section reduction techniques, typical radar cross sections are expected to be much smaller. If the radar cross section is less than the 10,000 square meters used in this example, the sensitivity requirement will be relaxed further. As a result – 80 dBmi would represent the maximum requirement on system sensitivity for ASM defence.

For the case of a navigation radar, the minimum detection sensitivity from Fig.9 for mainbeam intercept of the PILOT radar is – 85 dBmi and sidelobe illumination is – 115 dBmi at a free–space range of 35 km.

From the above discussion, current EW receivers do not have the sensitivity for the detection of current and projected LPI radar signals. A system sensitivity requirement of about – 100 dBmi should be adequate even for over the radar horizon operation. LPI operation is achieved by reducing the peak transmitted power while the duty cycle is increased. As a consequence, the expected signal level would be lowered by a factor of 30 dB. The future signal environment will consist of both LPI and conventional signals and thus additional dynamic range is required.

5.0 LARGE TIME–BANDWIDTH RADAR SIGNALS

Large time–bandwidth radar signals are used in a number of radar applications such as search–surveillance, tracking, ground mapping, radar imaging ,etc.[1]. A large time–bandwidth signal is usually generated either by frequency or phase modulation to widen the signal bandwidth [2,3].

5.1 Pulse Compression

Large time–bandwidth signals have been widely used for pulse compression applications. Pulse compression allows a radar to utilize a long pulse to achieve large radiated energy, but simultaneously to obtain the range resolution of a short pulse[3]. A radar designer can simultaneously increase detection range (average transmitted power) and maintain (or improve) the range resolution through pulse compression techniques without increasing peak power transmitted by the radar. There are also disadvantages associated with the use of pulse compression waveforms. If the leakage signal entering the receiver causes degradation to the performance of the receiver or is greater than the damaging level, the minimum range is set by the transmitted pulse width[3]. In addition, the waveform usually generates self–clutter and range sidelobes. Finally more complex receiver and transmitter signal generation and processing are required.

There are many waveforms which can be used for pulse compression applications. They include linear FM, nonlinear FM, discrete frequency shift, polyphase codes, Barker codes, maximum length sequences, compound Barker codes, code sequencing, complimentary codes, pulse burst,and stretch[3].

5.2 Linear FM on Pulse

Among the frequency modulation signals , the linear FM on pulse (LFMOP) has been more widely used. A scatter diagram on LFMOP radars is shown in Fig. 10 [1] which represents those already in the inventory as well as the more advanced radars under development. From Fig. 10, the general bounds on the LFMOP signals are : (a) time–bandwidth product (β) varies from 3 to 100,000 (b) pulse width (PW) from 250 ns to 2 ms and (c) bandwidth (B) from 60 KHz to 600 MHz. Most of the radars employ parameters of less than 10 MHz in bandwidth, pulse duration on the order of 10 to 100 μs and time–bandwidth product around 100.

The rate of change is usually a more convenient parameter to measure. The absolute rate of change of the LFMOP signals versus pulse width and bandwidth are plotted in Figs. 11 and 12 respectively. The absolute rate of change varies over 6 orders of magnitude from 1500 MHz/μs to 5x10^{-4} MHz/μs. From Fig. 11, the rate of change is approximately inversely proportional to $1/PW^2$ while from Fig. 12, it is proportional to B^2.

The scatter diagrams on LFMOP extracted from existing inventory are plotted in Figs. 13 to 15. The total number of signals are 47. As expected the ranges are smaller due to older technology. By comparing the two sets of plots, the changes have been for wider bandwidth, longer pulse duration and thus larger time–bandwidth products.

5.3 Phase Modulation on Pulse

Table I lists parameters for some phase modulation radars[1]. There are a number of signals with Barker code 13. The smallest bit length is 0.0125 μs which implies an instantaneous bandwidth of 80 MHz. The majority are less than 10 MHz. A similar survey has also been carried out from a data base and a very limited number of phase modulation radars are found. The majority are of Barker Codes 11 and 13.

5.4 Wideband Signals for LPI Applications

Most of the waveforms mentioned above have been designed and used for pulse compression applications. There are no reasons why they can not be adapted for LPI applications. As discussed in Section 2.2, for LPI applications the pulse duration has to be greatly increased while the peak power is reduced to keep the average transmitted power

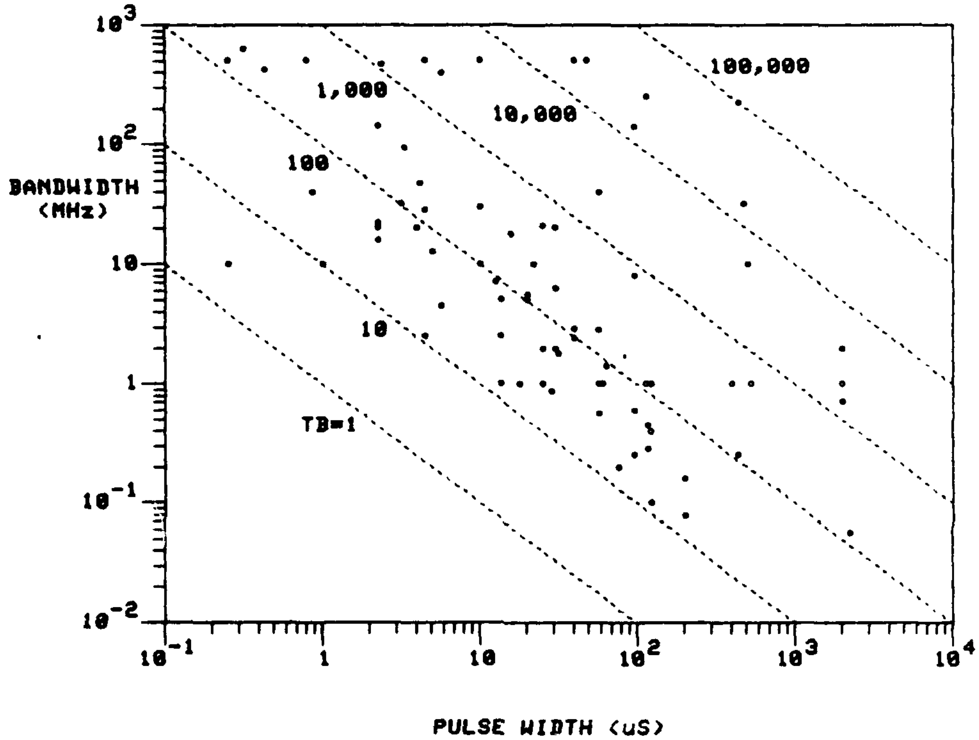

FIGURE 10. SCATTER DIAGRAM OF FREQUENCY DEVIATION VERSUS PULSE WIDTH FOR SOME LINEAR FMOP RADARS
(Source: " Electronic Intelligence: The Interception of Radar Signals", Richard G. Wiley, P.142)

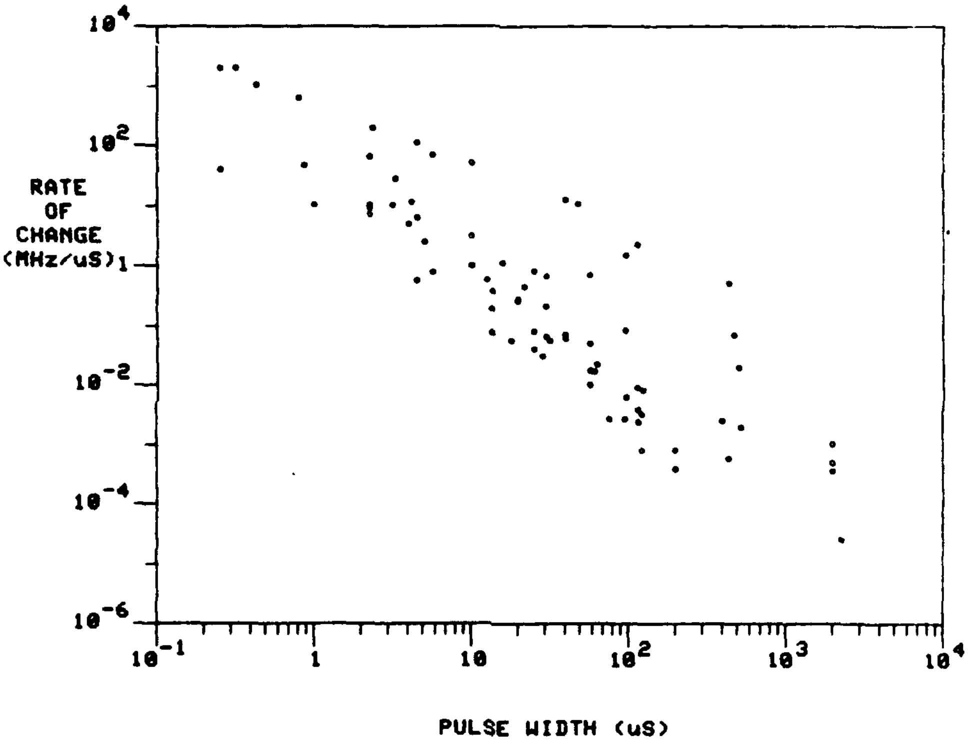

FIGURE 11. SCATTER DIAGRAM OF RATE OF CHANGE IN FREQUENCY VERSUS PULSE WIDTH FOR SOME LINEAR FMOP RADARS (Source: " Electronic Intelligence: The Interception of Radar Signals", Richard G. Wiley, P.142)

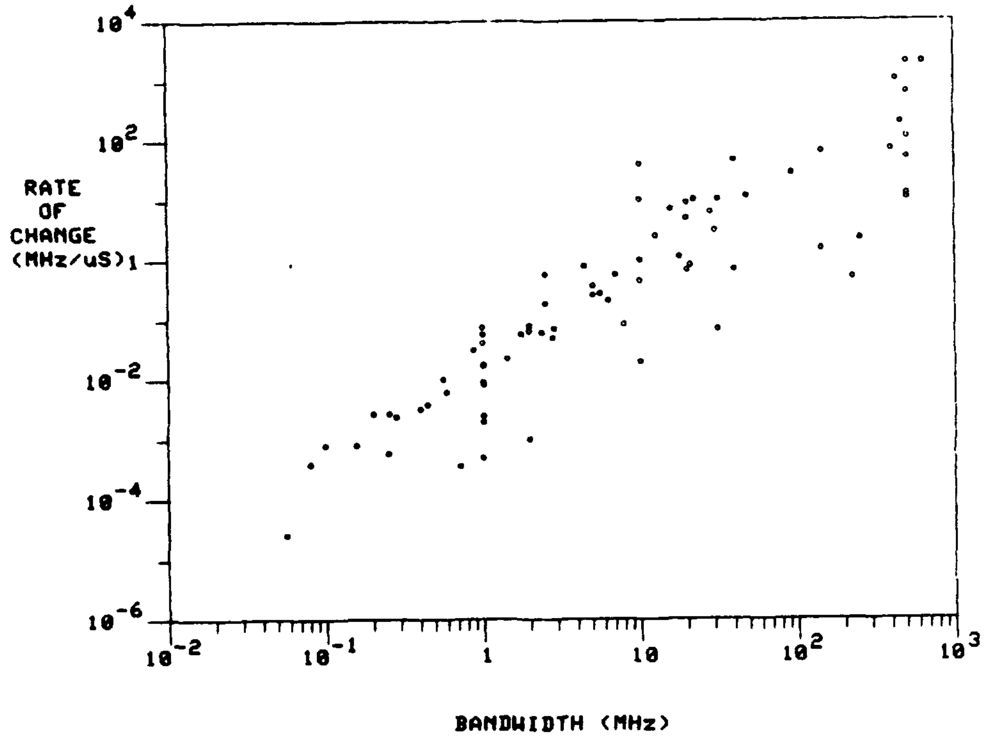

FIGURE 12. SCATTER DIAGRAM OF RATE OF CHANGE IN FREQUENCY VERSUS BANDWIDTH FOR SOME LINEAR FMOP RADARS (Source: " Electronic Intelligence: The Interception of Radar Signals", Richard G. Wiley, P.142)

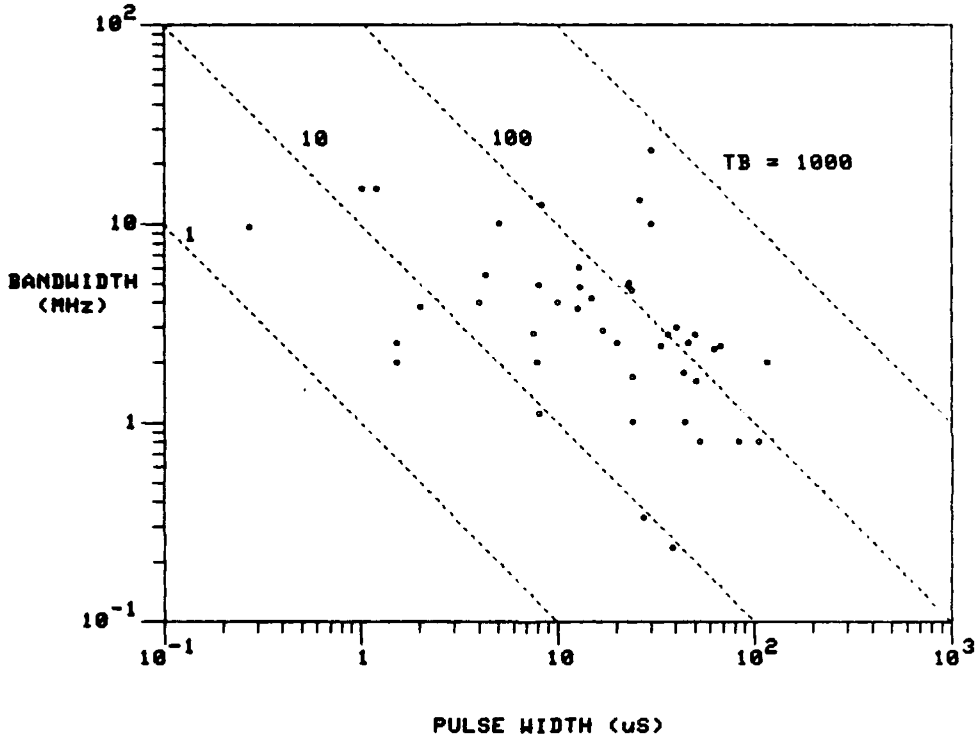

FIGURE 13. SCATTER DIAGRAM OF FREQUENCY DEVIATION VERSUS PULSE WIDTH FOR SOME LINEAR FMOP RADARS

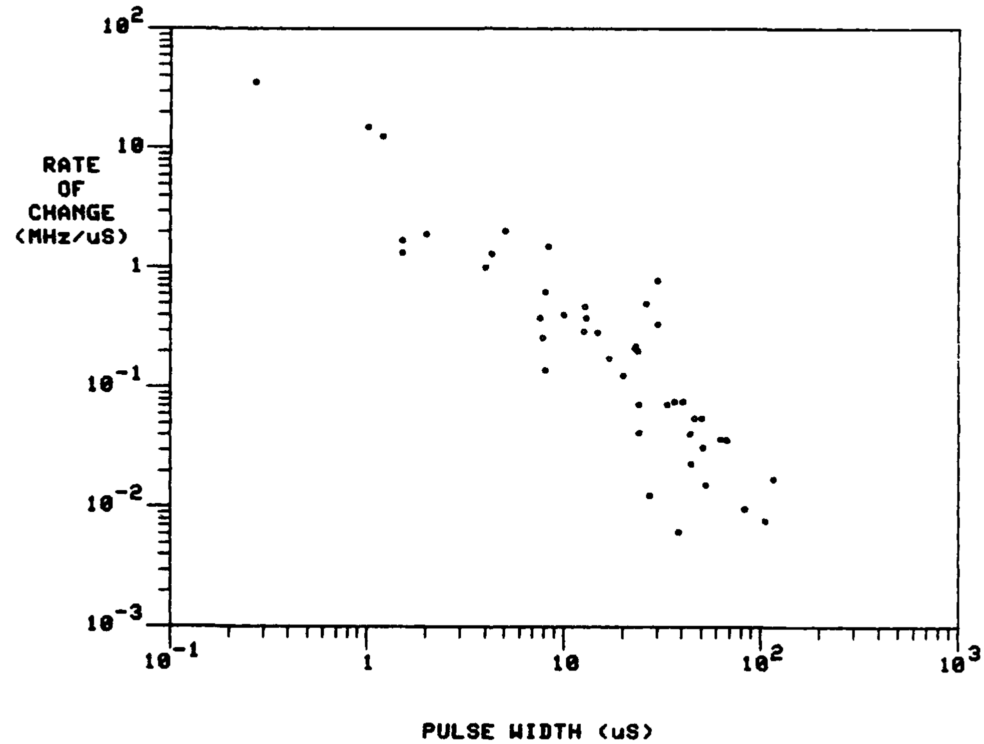

FIGURE 14. SCATTER DIAGRAM OF RATE OF CHANGE IN FREQUENCY VERSUS PULSE WIDTH FOR SOME LINEAR FMOP RADARS

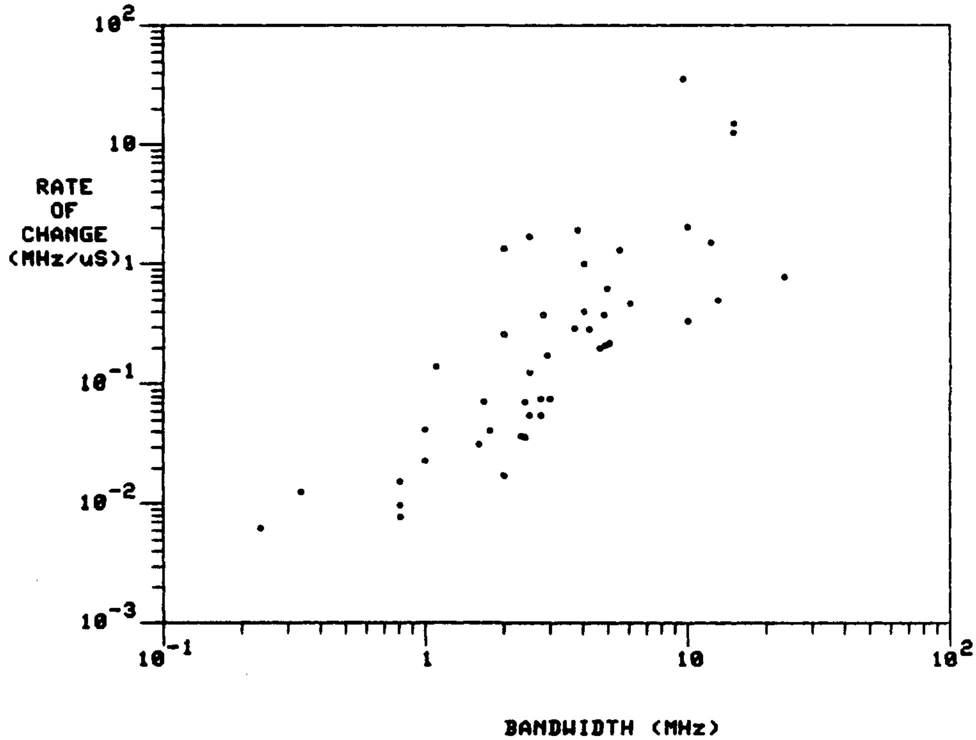

FIGURE 15. SCATTER DIAGRAM OF RATE OF CHANGE IN FREQUENCY VERSUS BANDWIDTH FOR SOME LINEAR FMOP RADARS

TABLE I

Parameters for Some Phase Modulation Radars

Phase Coding	RF (GHz)	Pulse Width (μs)	Bit Length (μs)
180° Reversal	0.20	2000	20.0
	0.42	6	0.46
	0.43	390	6.09
	—	780	6.09
	0.60	52	4.0
	5.6	0.2125	0.0125
	9.35	18.1	0.203
	16.25	2.0	0.154
	9.0	CW	125 (511 bits)
	9.0	CW	1.5 and 0.3 (63 and 255 bits)
	10.0	CW	2.19
	10.0	CW	0.0133 (8,000 bits)
	3.0	6.6	0.5
	5.5	6.0	0.1
	8.75	CW	Unknown (range resolution = 25 m)
	10.2	CW	Unknown (range resolution = 35 m)
Other			
Quadrature	—	25.6	0.2
Frank			
Polyphase	—	39.2	0.0625
	—	78.4	0.108

Source: " Electronic Intelligence: The Interception of Radar Signals "
Richard G. Wiley, P. 144

the same. The optimum waveform to be used greatly depends on the information to be extracted by the radar and is application dependent. Some radars may require waveforms which are more tolerant of doppler shifts while others may be used when jamming or EMC is a problem. Other important factors to be considered are the cost, weight and complexity.

6.0 LPI RADAR RECEIVING ARCHITECTURES

It has been shown that LPI signals are of much longer signal duration and are most likely on the order of 10 μs and more. Lower peak power is also used with either frequency or phase modulation. Large bandwidth signals greater than 10 MHz may not be needed unless very high range resolution is required. This implies that signals of relatively narrow bandwidths and high duty cycles are effective for LPI applications. In terms of receiver requirement, it is concluded from Section 4.2 that for LPI radar signal detection, a receiver system is required to have a sensitivity of approximately –100 dBmi and a correspondingly larger dynamic range.

LPI radar receiver makes use of the coherent integration gain of matched filters and incoherent integration gain by integrating a number of pulses. On the other hand, current EW receivers are designed to detect conventional radar pulses which are short in duration and over a broad frequency band. As a result, the radar receiver has an advantage over an EW receiver by the time–bandwidth factor (τB_i). This time–bandwidth factor is quite large for current EW receivers. A radar designer will try to maximize τ by employing LPI signal waveforms. There is no doubt that the EW receiver designer will respond by minimizing B_i to match these waveforms.

Current wide–open EW receivers such as the IFMR and crystal video receivers work well in a low density signal environment where the pulses are short in duration. However they are susceptible to interference in a dense signal environment where radar pulses may overlap in time. This problem has become more severe with the introduction of pulse compression waveforms and pulse dopplers which are higher in duty cycles. The problem associated with signal overlapping may become worse with LPI signals which are expected to be of even higher duty–cycles. On the other hand, LPI signals are expected to be of much lower in peak power and thus those LPI radars which are far away will not affect the performance of the EW receivers. However, there are likely " friendly " LPI radars on the same platform and nearby which will cause interference. As a result, with the proliferation of pulse compression and LPI signals, current wide–open IFM and crystal video receivers will be more susceptible to the problem of interference and thus are poor candidates for future ESM systems. In addition, they do not have the sensitivity for the detection of current and projected LPI signals and thus are not considered.

We will investigate some of the potential architectures which could be used to augment an existing shipborne radar ESM system. In the existing multi–band ESM system, the output of an omni antenna is fed to an IFM receiver for the determination of frequency. It is also fed to other crystal video detectors where other parameters such as PW, pulse amplitude and TOA are measured. DF is measured using an 8–port amplitude comparison system composed of crystal video receivers with RF preamplification. We will assume that the existing system will remain as the main ESM system as shown in Fig.16

and the improvement on the overall capability is carried out by adding LPI radar receivers. The auxiliary channelizer as shown in the figure is another add–on system which is used primarily to " see through " strong high–duty cycle signals located nearby.

There are three potential ESM architectures to be discussed in this report for the detection of LPI radars. These three potential architectures are by no means the only candidates for LPI signals detection. There are also other types of receivers such as the correlator and the fast–scan superhet as suggested by Wiley[1] which could be used for LPI signal detection. However due to the scope of this report, they are not considered here. An optimum architecture can be designed only after the specific requirement and the scenario the system is expected to operate in are given. In the following sections, we will discuss briefly some schemes which may have some potential in meeting the requirement for the detection of LPI signals.

6.1 Narrow–band Receiver With Feature Detector

As discussed in Section 2.2, the sensitivity of EW receivers can be greatly increased by using a narrower RF and video bandwidths. A good candidate is the channelized receiver as discussed in Section 3.0 (c) where a broad instantaneous frequency band is covered. The channelizer can either be implemented by a filter bank receiver , acousto–optic channelizer or microscan receivers. For discussion purposes, the filter bank or acousto–optic channelizer are used. If the video bandwidth is reduced to 0.1 MHz, the effective noise bandwidth (B_i) is 0.9 MHz and an improvement in sensitivity of 13.3 dB is obtained. In order to improve the sensitivity further, both the noise figure and transmission loss should be minimized on the omni–channel. If the noise figure can be reduced from 7 dB to 4 dB and the loss from 15 dB to 2 dB, then the system sensitivity will be – 97 dBmi with an output signal–to–noise ratio of 12 dB. The omnidirectional antenna has also been assumed to have a gain of 0 dBi.

A block diagram of this system architecture is shown in Fig. 16. The function of the channelized receiver is for LPI signal activity detection. A 500–MHz instantaneous bandwidth is obtained by using a 27–channel receiver. Once the frequency of the LPI signal is detected, the information can then be used to tune a set of eight identical superhet receivers to the eight–port network for bearing measurement. If the noise figure , loss, RF and video bandwidths of the superhet receiver in the DF channel are the same as the channelizer in the omni–channel, a maximum improvement in sensitivity of approximately 10 dB can be obtained due to the higher gain of the antenna in the DF channel. However the gain of the antenna decreases when the signal is received off boresight from the antenna. As a result, there may not be any improvement in terms of signal–to–noise ratio for bearing measurement using amplitude comparison unless the RF bandwidth of the superhet receivers is also reduced. Once the bearing of the signal is determined, a feature detector such as a fine frequency discriminator or a digital I/Q demodulator can be used to extract the modulation feature of the signal.

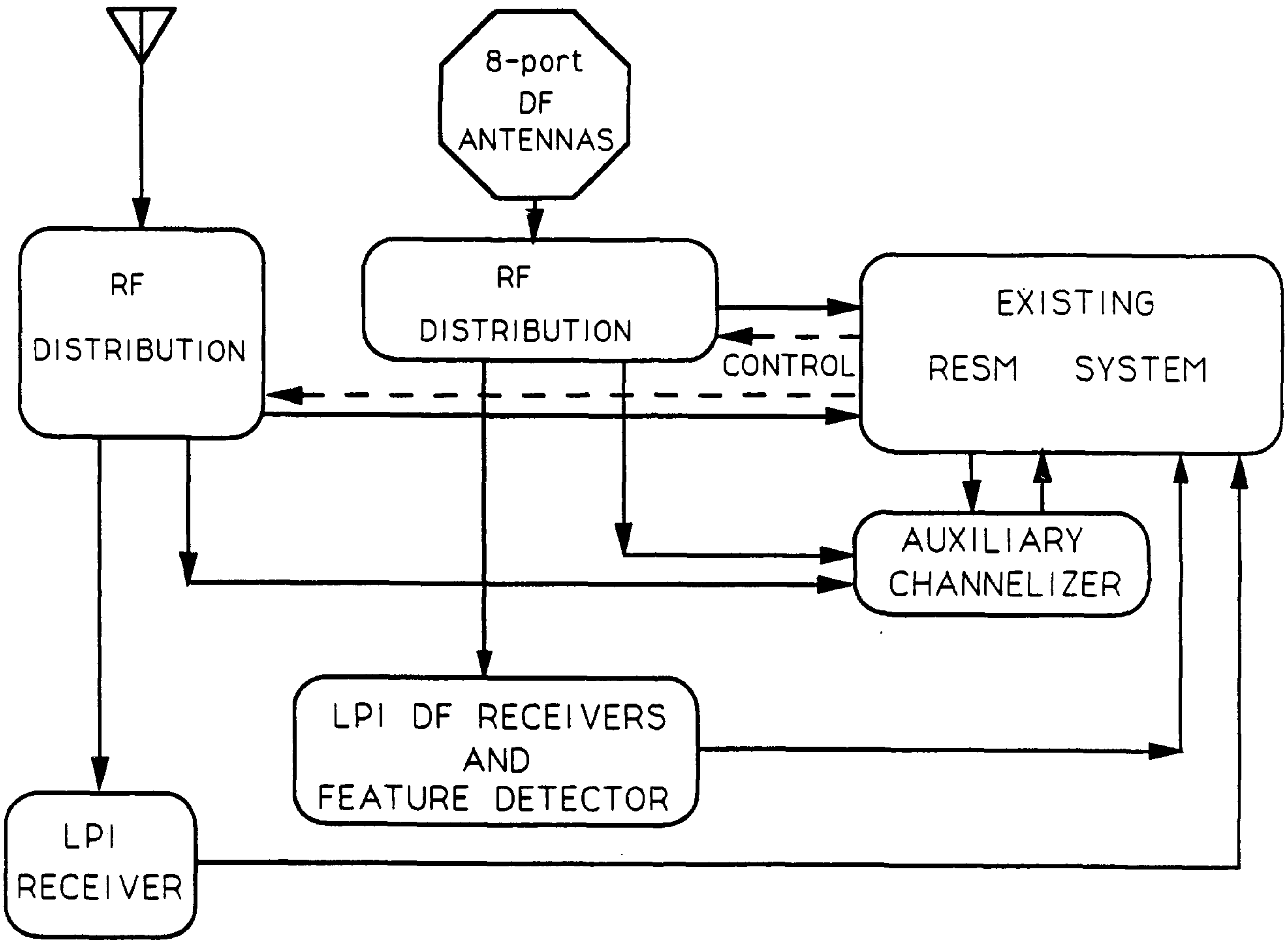

FIGURE 16. BLOCK DIAGRAM OF A LPI ESM RECEIVING SYSTEM

Using this system architecture with a narrow video bandwidth, the sensitivity of the system is improved for the detection of LPI signals. At the same time the response to conventional short pulses will also be greatly reduced. Ideally the width of the video filter should be matched to that of the LPI signal of interest. If the pulse width of a conventional pulsed signal is 0.1 μs and its peak power is 30 dB higher than a LPI signal of 10 μs in duration, the output power for the pulsed signal will be lowered by a factor of 20 dB as compared to the LPI signal. As a result, the output dynamic range requirement for detecting both conventional and LPI signals will be reduced.

If this system is used to detect the PILOT radar, it can detect the mainbeam with no difficulty. However, in the presence of other high power pulsed signals, it is still a very difficult task in picking out the LPI radar. One way is to compare the signals detected in the channelizer to those measured by the main ESM system. If a signal is not detected by the main ESM system but detected in the channelizer, then it is assumed to be a LPI signal. Another technique is to use a number of video filters attached to each channel with different video bandwidths. For example, if three video bandwidths of 1 , 0.1 and 0.01 MHz are used to detect a LPI signal of a minimum duration of 100 μs among a number of conventional pulsed signals with a maximum pulse width of 1 μs. Assuming the signal is located at the center of a channel and for the LPI signal, the signal–to–noise ratio from the three video detectors increases by approximately a 5 dB step as the video bandwidth is reduced by a factor of ten. On the hand, the output signal–to–noise ratio decreases by approximately a 5 dB step for a conventional radar pulse. A relative comparison of the three output levels will give some indication on whether the signal is of long or short duration and thus indirectly giving the nature of the signal. A final confirmation is carried out by the use of the feature detector. The video filters can either be implemented directly using analog video filters in parallel or by digitizing the output from the video detector and then digitally filter the data with three different video bandwidths. Another possible technique is just limit the input power signal level to the channelizer by the use of frequency selective limiters or just regular limiters so that the signal level of the strong pulsed signals will be greatly attenuated. Depending on the type of limiter used, various intermodulation products can be generated. The idea behind this scheme is first to limit the peak power of conventional pulsed signals to the channelizer, so that the differential power ratio of conventional strong pulsed signals to LPI signals will be greatly reduced. The use of a narrow video filter will further reduce the output signal of a conventional pulse to below the threshold.

Another processing technique to discriminate conventional pulses from LPI signals as suggested by Wiley[1] is to exploit the property that a LPI signal is of much longer in duration than a conventional pulsed signal. The output from each channel of a channelizer is detected and sampled. A count is declared once the threshold is crossed. A long duration signal will give more consecutive counts while the number of consecutive counts will be very small for short duration pulsed signals.

ι.2 Acousto–optic Receiver With Feature Detector

The receiver requirement of having relatively a large number of narrow channels with a narrow video bandwidth for the detection of LPI signals can easily be met by the use of a time–integrating acousto–optic receiver. The narrow video bandwidth and the relatively large number of channels can be implemented relatively easily by using a time–integrating photodetector array.

Considerable progress has been made recently on the development of both 1–D and 2–D acousto–optic receivers. In a 1–D configuration, the acousto–optic receiver performs spectrum analysis on the received radar signals while in the 2–D configuration both spectrum analysis and direction–finding are carried out [9]. An acousto–optic receiver which is suitable for the detection of LPI signals can be implemented easily using " off–the–shelf " photodetector arrays with variable integration times.

An analysis on the sensitivity of both the 1–D and 2–D acousto–optic receivers have been given[10] and the results are summarized as follows.

The processing gain is defined as the ratio of the input signal–to–noise ratio without integration to the input signal–to–noise ratio with integration in order to achieve the same detection probability. The maximum incoherent processing gain (PG) for a CW signal in a time–integrating acousto–optic receiver when the signal is located at the center of a photodetector element is given by

$$PG = \sqrt{B_s T\ J} \tag{29}$$

The input sensitivity of an acousto–optic receiver in dBm for an output signal–to–noise ratio = 1 is given by

$$\text{Sensitivity} \approx f(SNR_z) + KT_rB_s + F_i - \sqrt{B_sT_iJ} - N \tag{30}$$

where $f(SNR_z)$ is the sensitivity loss factor which is a function of the input receiver noise, detector noise and quantization noise[10], B_s is the noise equivalent bandwidth, J is the number of samples integrated. For an acousto–optic spectrum analyzer, N = 1, For a 2–D acousto–optic receiver N = number of Bragg cell channels or antenna elements.

For a non–CW signal, the sensitivity will be degraded by a mismatch factor given by

$$\text{Mismatch Factor} \approx T_1/T\ \mathscr{H}(f) \tag{31}$$

where T_1/T is the total duration of the signal intercepted to the integration period of the photodetector array , $\mathscr{H}(f)$ is the spectral factor normalized by the spectral distribution of a CW signal. Both T_1/T and $\mathscr{H}(f)$ are less or equal to 1.

In order to maximize the input receiver sensitivity, the receiver parameters must be designed to match the signal of interest in terms of matching the integration period to the duration of the signal and the frequency bin size to the total spectral width of the signal.

An acousto–optic receiver operating in an integrating mode is essentially an energy detector. As discussed in Section 2.0, a LPI radar will transmit a much lower peak power than a conventional radar. However the total energy or average power will be about the same. As a result, the output power levels from the acousto–optic receiver will be about the same when both types of radars are intercepted.

6.2.1 Acousto–optic Spectrum Analyzer

The acousto–optic receiver is assumed to have the following parameters:

F_i	4 dB
L_i	2 dB
G_i	0 dBi for Omni antenna, and
	10 dBi for DF antenna
J	1 sample integrated
B_s	0.5 MHz
T	0.25 msec integration period, and
$f(SNR_z)$	2 dB

Using Eq.(30), the CW sensitivity of the acousto–optic receiver with the above parameters is calculated to be –121.5 dBm for S/N = 0 dB and the system sensitivity is –119.5 dBmi.

When the PILOT radar is operating at its maximum search range, the FMCW frequency sweep is 1.5625 MHz. If this acousto–optic receiver is used to intercept the PILOT radar at this maximum search range, some degradation in sensitivity will occur and the factor is approximately 0.3 or 5 dB. Therefore the net system sensitivity is –114.5 dBmi. Assuming $(S/N)_i$ equals 12 dB, the system sensitivity is reduced to – 102.5 dBmi which is more than adequate for the mainbeam interception of PILOT.

Since the acousto–optic spectrum analyzer is an energy detector, the other pulsed signals will also be detected. For the PILOT radar, the acousto–optic spectrum analyzer can discriminate it by monitoring the detected peak frequency location of the signal over a number of integration periods. The peak location of the signal will move back and forth with the frequency rate and span of the PILOT signal.

When the PILOT is operating at other modes for the shorter range of operation, the sweep in frequency will be larger. This will cause some degradation to the sensitivity of the acousto–optic receiver to this type of signal. However, the discrimination of the PILOT radar from other conventional radars using the above two methods will be easier.

The effective integration time (video bandwidth) of the acousto–optic spectrum analyzer can be adjusted to match the duration of the signal intercepted for maximum sensitivity. This can be accomplished easily by either changing the integration period on the photodetector array or changing the number of samples integrated digitally.

A block diagram of the potential LPI ESM system is shown in Fig. 16, where the acousto–optic spectrum analyzer is employed in the omni–channel. It is used as a LPI signal activity detector and coarse frequency information can also be obtained. Once the frequency of the signal is measured, a set of eight narrow–band superhet receivers as discussed in Section 6.1 can be used to measure the bearing. The fine frequency structure of the signal can be extracted by the use of either a frequency discriminator or a digital I/Q demodulator.

6.2.2 2–D Acousto–optic Receiver

For the 2–D acousto–optic receiver, a separate antenna array and down–converter are used. The following parameters are assumed for the analysis:

F_i	4 dB
L_i	2 dB
G_i	0 dBi for each 2–D antenna, and
	10 dBi for each DF antenna
J	1 sample integrated
B_s	2 MHz
T	10 msec integration period, and
$f(SNR_z)$	2 dB
N	5 channels

Following a similar analysis as given in Section 6.2.1, the receiver sensitivity for CW signal is – 133.5 dBm and system sensitivity is –131.5 dBmi for S/N = 0 dB.

When the PILOT radar is operating at its maximum search range, the FMCW frequency sweep is 1.5625 MHz. If this receiver is used to intercept the PILOT radar at this maximum search range, the total illumination time (T_I) is 3 msec (three sweeps). The mismatch factor will reduce the receiver sensitivity from –131.5 dBmi to –125.5 dBmi at S/N = 0 dB. For $(S/N)_i$ equals 12 dB , the system sensitivity is – 113.5 dBmi.

A block diagram of the LPI ESM system using the 2–D acousto–optic receiver is shown in Fig. 17. At a system sensitivity of – 113.5 dBmi, the mainbeam and the sidelobes (at a free–space range of 30 km) of the PILOT radar is detected by the 2–D acousto–optic receiver. The frequency and bearing information is then passed on to select one of the DF antennas and to down–convert the signal to either a narrow–band frequency discriminator or a digital I/Q demodulator for extracting the fine signal modulation. If the narrow–band frequency discriminator and digital I/Q demodulator are assumed to have a RF bandwidth of 20 MHz and a video bandwidth of 0.1 MHz, the maximum system sensitivity when the signal is received off boresight from the antenna is computed to be – 107 dBmi. The directional DF antenna will provide some spatial filtering and reduce some of the effects of multi–path while the 20 MHz bandwidth will provide some frequency selectivity.

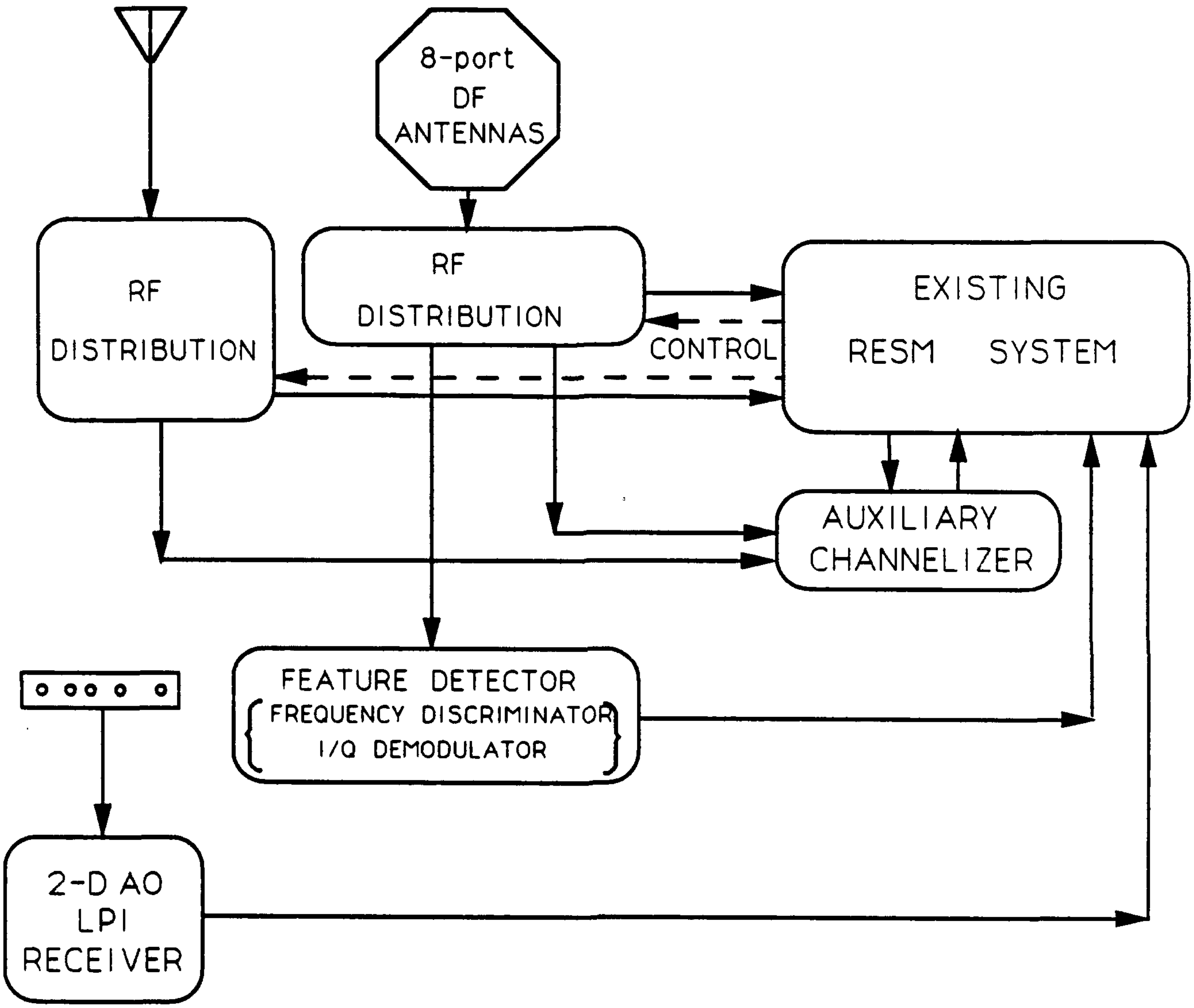

FIGURE 17. BLOCK DIAGRAM OF A LPI ESM RECEIVING SYSTEM USING A 2–D ACOUSTO–OPTIC RECEIVER

The special feature of the 2–D acousto–optic receiver over the acousto–optic spectrum analyzer is that the bearing as well as the spectrum of the signal are measured. Both the conventional and LPI signals will also appear in the output unless some limiting is carried out in the front end to limit the high power pulsed signals. When a long weak LPI signal and a strong short interfering signal are applied to a limiting amplifier, intermodulation products will result. The LPI signal will also be interrupted and suppressed during the presence of the strong signal. From the measurement done on a 2–D acousto–optic receiver, the bearing of the LPI is found to be slightly affected by the presence of the strong signal. This is due to the fact that all the channels are matched and the phase of the signal in each channel is affected by the same amount. As a result, the original phase differences among the channels are preserved and thus not affecting the resultant bearing measurement. The LPI signal will be suppressed during the presence of the interfering signal and the spectrum will be slightly affected as well. The degradation in terms of sensitivity loss will depend on the amount of time the LPI signal is interrupted and the effect should be small. More work is needed to exploit this technique for rejecting strong pulsed signals.

7.0 SUMMARY AND CONCLUSIONS

A radar receiver is designed to exploit the coherent integration gain of matched filters and the incoherent integration gain by integrating a number of pulses. On the other hand, current EW receivers are designed to cover a much broader RF bandwidth and to detect the shortest anticipated radar pulses and thus the resultant equivalent noise bandwidth (B_i) can be quite large. As a consequence, there is a mismatch between the radar transmitter waveform and the EW receiver. The relative mismatch is given by the time–bandwidth factor (τB_i) and it is quite large for some current wide–open EW receivers. Despite this mismatch, the EW receiver has the range advantage due to one–way propagation loss. In addition, most current radars transmit short duration pulses with relatively high peak power. As a result, most current radars can be detected easily by the use of current EW receivers.

To make a radar LPI in which the radar cannot be intercepted beyond the range at which it can detect targets itself, a radar designer can maximize the mismatch further by increasing the duration (τ) of the signal. This can be carried out by employing signal waveforms in which the range resolution of the radar is recovered while the transmitted peak power can be reduced. As a result, LPI signals are expected to be of long duration and thus higher duty cycles. The EW receiver designer can also respond by minimizing B_i to match these LPI waveforms. However, it is difficult to build an EW receiver which can meet both the requirements of having a small equivalent noise bandwidth and be able to detect signals over a wide instantaneous RF bandwidth.

Three radar functions, namely search, ASM missile seeker and navigation, have been examined against current EW receivers for LPI operation. It has been shown that LPI operation is most easily achieved at close ranges only. In the search function, the range is usually quite large and the target size can be small. As a result, it is very difficult to design a radar LPI against conventional EW receivers when the mainbeam is intercepted. A combination of antenna sidelobe control and waveform coding are essential for LPI operation when the interceptor is located in the sidelobes. For the ASM seeker function the target size is relatively larger and the range is reduced when tracking a closing target. As a result, the techniques of power control and waveform coding can be effective for LPI operation. However, the complexity ,cost and space will probably limit their use in practice until technology improves in the future. For the function of navigation, the range is relatively short and there are already LPI radars in operation such as the PILOT which makes use of waveform coding. However, no matter which LPI technique is used the introduction of radar cross section reduction techniques will make LPI operation less effective.

The general system sensitivity requirement for the detection of current and projected LPI radars is found to be on the order of – 100 dBmi which cannot be met by current EW receivers. However with some modification to current narrow–band EW channelizers in terms of reduced video bandwidth, the sensitivity can be improved for LPI radar detection.

Three general LPI ESM architectures, using narrow–band channelizers, superhet and acousto–optic receivers, have been examined in this report for shipborne applications. They have shown some promise in terms of providing the sensitivity and capability in an environment where both conventional and LPI signals are present.

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(U) Most current radars are designed to transmit short duration pulses with relatively high peak power. These radars can be detected easily by the use of relatively modest EW intercept receivers. There radar functions, namely search, anti-ship missile (ASM) seeker and navigation, are examined in this report to evaluate the effectiveness of potential low probability of intercept (LPI) techniques, such as waveform coding, antenna profile control and power management, that a radar may employ against current EW receivers. The general conclusion is that it is possible to design a LPI radar which is effective against current intercept EW receivers. LPI opeation is most easily achieved at close ranges and against a target with a large radar cross section. The general system sensitivity requirement for the detection of current and projected LPI radars is found to be on the order of - 100 dBmi which cannot be met by current EW receivers. Finally, three potential LPI receiver architectures, using channelized, superhet and acousto-optic receivers with narrow RF and video bandwidths are discussed. They have shown some potential in terms of providing the sensitivity and capability in an environment where both conventional and LPI signals are present.

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